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CONTROL TRANSLATION SERIES

Measurement Techniques

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SOVIET INSTRUMENTATION AND CONTROL TRANSLATION SERIES

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Measurement Techniques

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IMPROVEMENT OF THE QUALITY OF PRODUCTION IN INSTRUMENT-MAKING

B. L. Sokolov

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 1-3, March, 1961

The enormous tasks outlined by the 21st Party Congress, the June (1959) and July (1960) plenary sessions of the CPSU (Communist Party of the Soviet Union) Central Committee, on speeding up technical progress, introducing new equipment, production automation, and raising labor productivity, cannot be carried out without the wide application of modern high-quality measuring instruments in our national economy.

The growth of instrument-making places important duties on the Committee's local agencies, which should take the necessary measures for improving the quality of production, discontinuing the manufacture of obsolete instruments, carrying out state and routine testing of the manufactured instruments on a high technical level, etc. In this connection the personnel of the Committee's local agencies must intensify their control of the instrument-making industry.

The GKLs (State Inspection Laboratories) should establish closer relations with the planning authorities of the Sovnarkhozes (Councils of National Economy) and regional Soviet executive committees, acquaint themselves with the planning of instrument production, with the dates set for completing the development work, for making experimental models and starting their mass production. It is important to insist that the State Planning Committees of Union Republics and the Sovnarkhozes should plan new instrument production and should transfer the manufacture of instruments from one establishment to another by agreement with and taking into consideration the opinions of the Committee's accredited representatives at the Councils of Ministers of the Union Republics and of the heads of the GKLs. The provision should be strictly enforced that the mass production of a new instrument can only be planned when it is known that this instrument has successfully passed state testing, which should constitute one of the stages in the development of the instrument.

Close relations should also be established with scientific research institutes and design offices which are developing new types of instruments. A routine should be established to make the production even of a single type of instrument impossible without the knowledge of the GKLs and the committee.

As the result of their business relations with the planning authorities the GKLs should have precise plans for the development and production of new types of instruments, improvements and modernization of the equipment in production, discontinuing the manufacture of obsolete instruments, and their replacement by modern designs. It is necessary systematically to check through the agency of the Sovnarkhozes and Soviet regional executive committees the assortment of the instruments being produced.

The heads of the GKLs should take part in the regional and republican conferences, raising there the problem of improving planned production of new instruments and the specialization of plants, and strengthening the methodical supervision of instrument-making by the local scientific research organizations.

In some instances directors of instrument-making plants tolerate infringements of government regulations on the procedure in testing and manufacture of instruments and in their state testing, as well as the infringement of the Committee's regulation 12-58 and 2-59. The GKLs should keep an up-to-date list of establishments which are producing or intend to produce measuring instruments, and demand that they adhere to the Committee's regulations. Organizational and technical instruction of the personnel of instrument-making plants should be carried out at the proper time on all the problems involved in the production of measuring instruments.

One of the methods by which the GKLs can actively effect an improvement in the work of instrument-making plants consists in the participation of the GKLs in the factory technical conferences on the quality of production, in technical commissions, workshop meetings and in factory party conferences. The GKLs should systematically participate in the technical conferences organized by the chief engineers of instrument-making plants for discussing the results of routine instrument testing. The GKLs should make proposals at these conferences for practical measures to be taken in order to eliminate defects discovered in testing. Judging by the experience of the Ivanovo GKL such conferences produce good results, make it possible to eliminate defects rapidly, and efficiently to solve organizational problems.

It is advisable to hold conferences at the instrument-making plants on the quality of production, with the participation of consumers' representatives, and to establish the practice of periodic check dismantling of instruments,

thus making it possible to discover directly any defects and to take the required measures for their elimination. It is also important to hold technical conferences on the results of the state testing of instruments. Such conferences are held by the personnel of the Sverdlovsk branch of the VNIIM (All-Union Scientific Research Institute of Metrology) in conjunction with the local GKLs on the premises of the inspected plants.

Every GKL must assist the instrument-making plants in improving their production. For this purpose they should take part in working out measures for the elimination of defects discovered in routine testing, in studying the instruments' operational qualities, and in analyzing complaints. The GKLs should also present recommendations on improvement and modernization of instruments in production, on starting the production of improved types of instruments, on the provision of reference instruments, and on organizing checking and service inspection of the measuring equipment. The GKLs should daily check the correction of defects in instruments within the specified time, the extent and quality of factory testing of instruments, the efficient and timely preparation for state checking of instruments which are ready for delivery.

The GKLs must see to it that the manufactured instruments are up to the modern level of technical development and meet the requirements of our national economy. They should carry out this work in conjunction with the design office personnel and leading technical workers in the plants and Sovnarkhozes.

It is necessary to attract the personnel of factory test laboratories, scientific research institutes, design offices, etc., to the work of improving the manufacture of measures and measuring instruments, and of organizing the systematic exchange of experience between factory test laboratories (conferences, information on instrument-making problems, demonstration at the plants of the most modern means and methods of measurement, etc.).

According to the experience of the Krasnodar and Ivanovo GKLs and others, it seems advisable to allocate GKL personnel to each instrument-making establishment for supervising all the measures taken in the production of instruments and for checking their execution. This makes it possible to study any given enterprise, to analyze thoroughly the reasons for rejects in production, to analyze them systematically, to conduct efficiently surprise inspections, to check the observance of technological procedures, and to take any measures connected with improving the quality of production.

In the plants where the finished products are checked by state inspectors, the GKLs should establish control and checking stations (KPP), whose operations should be based on "Regulation on control and checking stations", approved by the head of the GKL in consultation with the factory administration.

The tasks of the KPPs are not limited to state testing of finished instruments. The KPP personnel should also study, together with the factory personnel, the design and operational properties of the instruments; they should make a systematic study of complaints received by the plant, and together with the factory personnel take the required measures for correcting defects in the instruments; they should make routine tests of the manufactured instruments, supervise new development and modernization of instruments, make proposals for discontinuing the production of obsolete instruments and starting the production of new improved instruments. The KPP personnel should also report to the head of the GKL any infringements or non-observance of specifications regarding the quality and the method of production of instruments, in order to take the required steps for correcting these infringements, they should check the condition of all the measuring equipment at the plant and take the necessary measures for its improvement.

The heads of the KPPs should not tolerate any deviation from specifications, GOSTs (All-Union State Standards), or the established manufacturing procedure.

The GKLs should pay special attention to the condition and quality of technical information compiled by the designers and submitted to the Committee's institutes together with the instrument models subject to state testing. The lack of attention on the part of GKLs to this question often leads to instructions based on obsolete standards, contrary to existing standards and lagging behind the latest developments in instrument-making.

It is known that the study of the operational properties of instruments helps to discover latent defects and improve the quality of the instruments. Such studies should be combined with routine testing and conducted by the factory personnel. Valuable experience has been gained by the Leningrad Sovnarkhoz, which has organized a "Laboratory of Reliability", where work is being conducted with the aim of increasing the life of the instruments and improving their quality. The Krasnodar GKL, which was instrumental in bringing about a considerable improvement in the quality of the instruments produced in its region, is not content with analyzing rejects and studying complaints received by the factories. It sends questionnaires on the quality of the instruments to every region of the Soviet Union.

Attention should be paid to raising the role of the GKLs, the design offices and instrument-making plants in introducing new measurement equipment into our national economy. It is known that the instrument-making plants are obliged to master the production of new improved instruments in a given time. In many instances the factories delay the development and adoption of new types of instruments. The GKLs should insist through the leading regional and territorial organizations that the development of new types of instruments be completed in time, plans for the

adoption of new designs be put into effect, the production of obsolete measurement equipment be forbidden, and should stop if necessary the state testing of defective or obsolete instruments.

One of the basic activities of the GKLs which require an improved and efficient organization is the routine testing of instruments. Many defects exist to date in these tests, such as the failure to combine these tests with a study of the operational qualities of instruments, and a negligent attitude to the enforcement of technological processes, to checking the state of the production equipment and to the work of auxiliary workshops and subsidiary plants. The defects of routine testing also include the lack of the enforcement by the GKLs of the GOSTs and the technical regulations on the packing of instruments and the complete supply of specified components. In many instances the GKLs have been guided in their work by obsolete GOSTs or have not been aware that the tested instruments have been standardized.

At present routine testing of instruments has grown in importance. These tests are the main method used by the Committee's agencies for state-testing the quality of instrument-production. It is, therefore, necessary to change in many respects both the form and technique of these tests.

It is necessary to involve in them on a wide scale the experts of the purchasing organizations, thus making it possible to discover certain defects in the instruments which it would have been difficult for the GKL personnel to find, and to provide tests with respect to parameters which would have been omitted by the GKLs and the manufacturing plants owing to the lack of the required devices, equipment and test conditions.

Routine testing is most efficient when a group of instruments similar to each other in their application and design and having standardized units and components are tested at the same time. This provides not only a saving in time but also makes it possible to arrive at conclusions regarding the quality of production and the observance of technical requirements in several types of instruments simultaneously. Defects discovered in one group of instruments make it possible to evaluate other groups if in their design similar circuits, units and components are used.

Great attention should be paid to the planning of routine tests, which should be systematically reviewed. The testing procedure should be arranged in such a manner as to reveal whether the instruments attain modern technological standards and satisfy the requirements of our national economy. The tests should also reveal any defects in the instruments and the reasons for these defects. The testing procedure should be determined by the experts of the organizations taking part in the tests.

The manufacturing quality of the instruments should be determined first of all by examining the customers' complaints and studying the operational qualities of the instruments as well as the results of factory testing of these instruments. Next, sample instruments should be tested to ascertain whether they meet the specifications, GOSTs and instructions of the Committee. The instruments should also be tested with respect to the parameters whose shortcomings were found in studying the operational qualities of instruments or were contained in the customers' complaints.

The GKLs should examine all complaints carefully and should not be satisfied with the usual formal answer that the defect was due to unsatisfactory transport conditions.

If the methods for testing certain parameters are not provided for by the GOSTs and by the Committee's instructions, the GKLs should be guided by the technical instructions in agreement with the appropriate institute of the Committee.

The GKLs should considerably improve their work in checking the fulfilment of the suggestions made by various establishments as a result of their testings. The implementation of the suggestions should be checked at the specified time instead of waiting for the time of routine testing.

If it is found that an instrument is unsuitable or obsolete it is necessary to send to the Committee a substantiated request for prohibiting the production of such instruments.

It is desirable for the GKL personnel engaged in supervising instrument-making to acquaint themselves with the entire technological process in the design and production of instruments.

The personnel of GKLs on whose territory there have been in the past no instrument-making plants should be trained in instrument inspection at other laboratories.

The Committee and its republican administrations should periodically hold seminars for the GKL personnel, acquainting them with the latest achievements in instrument-making and measurement techniques, and organize special conferences for various branches of instrument-making. It is also necessary for the GKLs to receive up-to-date and exhaustive information on the condition of instrument-making both at home and abroad.

The workers of the Committee's system have all the possibilities at their disposal to ensure that our industry produces in the near future measuring instruments only of the highest quality.

LINEAR MEASUREMENTS

NEW DEVELOPMENTS IN COORDINATE MEASUREMENTS

T. S. Gladilina

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 4-7, March, 1961

The coordinates of hole centers in plates and bridges of watches and other instruments are measured on universal measuring microscopes or on coordinate boring machines. These instruments and machines are inefficient for measuring coordinates, since they are not intended for such measurements.

This is why our industry has developed to a specification supplied by the Horological Scientific Research Institute a coordinate measuring microscope MKI intended for accurate and productive measurements of coordinates in miniature details, as well as for the measurement of distances and angles.

The maximum lengths measured by the MKI in two mutually perpendicular directions are 100 × 100 mm, the microscope graduations for measuring length amount to 1μ and for angles to 10"; the error in length measurements amounts to 2μ.

The appearance of the instrument is shown in Fig. 1. The longitudinal table 2 and the transverse carriage 3 can be displaced on the baseplate 1 in two mutually perpendicular directions by means of a rack-and-pinion drive. The millimeter glass scales,which have 100 divisions at intervals of 1 mm,are attached to them and displaced with them. For a rough adjustment the table and the carriage are displaced by fly-wheel drives 4 and 5 (drive 5 is located behind knob 7 and cannot be seen in Fig. 1). Their accurate displacement is made by means of knobs 6 and 7. Their linear displacements are read off the scales (Fig. 2) in the field of vision of the measuring microscope 8 with a magnification of 50, mounted on the baseplate. The scale with eleven pairs of bisection lines and a vertical index line is fixed, and the one with 100 graduations is adjustable by means of the fly-wheel drive 9. The figures inscribed above the bisection lines denote tenths of a millimeter. Calibrations of the millimeter scale serve as pointers for reading off the above scale. The figures on the moving scale denote hundredths of a millimeter, and each division of this scale read against the fixed index denotes microns. The optical circuit of the instrument is arranged in such a manner that the longitudinal and transverse millimeter scales are seen alternately in the microscope's field of vision. These scales can be seen through one microscope owing to their alternate illumination by means of the control board 10 and a special system of prisms. Moreover, the field of vision is illuminated by a yellow-green color for the longitudinal, and a blue-green color for the transverse scale.

The vertical guides of the transverse carriage serve to displace a 75 mm tube 11 with a central microscope 12, which has an exchangeable eyepiece head. One head is linear with dotted crosshairs, another consists of 25 concentric circumferences with continuous crosshairs, and the third has a split (double) image. The linear head is intended for checking the instrument and umpire measurements, and the concentric head is for rapid sighting of through-hole centers. The microscope's field of vision for an amplification of 30 is 5.7 mm, and for an amplification of 50 it is 3.4 mm. The brightness of the field of vision illumination is controlled by means of the iris diaphragm 13 fitted with a calibration scale. For the purpose of illuminating the object by means of reflected light,illuminator 14 is used, which is fixed to the lower part of the microscope tube.

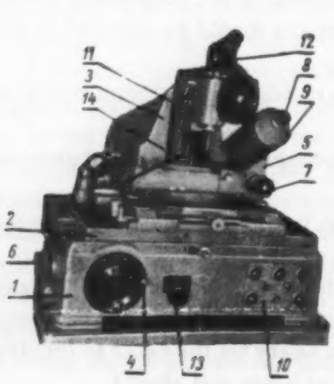

Fig. 1

Fig. 2

Fig. 3

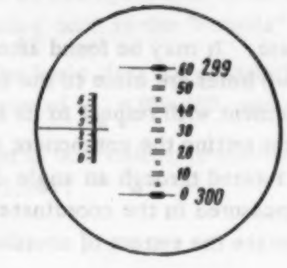

Fig. 4

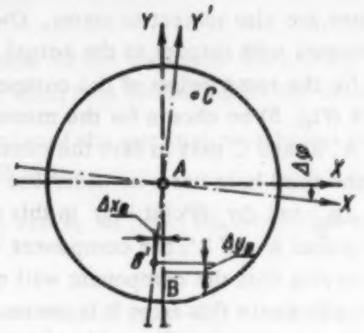

Fig. 5

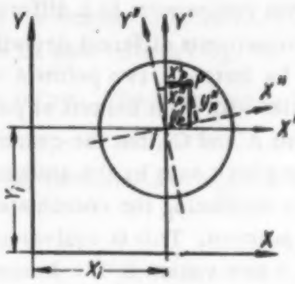

Fig. 6

The longitudinal carriage of the instrument is fixed to a plane table which is provided with an independent coordinate displacement with respect to the carriage by means of micrometer screws. The upper plate of the plane table carries a turntable 1 (Fig. 3) with a centering plate 2, which can also be displaced in two mutually perpendicular directions with respect to the turntable by means of screws. The angles of rotation of the table are read off an optical scale by means of microscope 3, whose field of vision is shown in Fig. 4.

The stages of the turntable and the centering plate are marked with crosshairs. Owing to the crosshairs and micrometer screws of the MKI instrument its stage and microscope can be rapidly centered and the object placed into position for measuring its coordinates.

The direction of coordinates in this microscope is the same as that adopted in mathematics and drawings, i.e., the longitudinal X-axis runs from left to right, and the transverse Y-axis runs from bottom to top, or away from the observer who stands facing the instrument.

Tests have shown that the productivity in measuring coordinates with the MKI instrument is 2-3 times higher than that attained with the UIM-21 microscope or with coordinate boring machines.

The articles are passed or rejected by determining the coordinate deviations Δx and Δy from the nominal dimensions of the centers of their holes or some other points:

$$\Delta x = x - x_0; \quad \Delta y = y - y_0, \tag{1}$$

where x, y and x_0, y_0 are the measured and nominal coordinate points respectively.

When the components are checked for deviation of the spacing between their axes the measurement results are processed by the graphic-analytical method [1].

Here we shall examine the processing of the measurement results when the position of the points is checked by the tolerances in their nominal coordinates.

The simplest and commonest method of processing these results is by means of a parallel displacement of the origin of coordinates along both axes. If the coordinate tolerances for all the checked points are the same the origin of coordinates is displaced independently along each axis X and Y by the amounts a and b respectively, in such a manner that the maximum plus or minus deviations from the new axes X' and Y' are the same; the coordinates of the new origin are determined from (2):

$$a = \frac{x_{max} + x_{min}}{2}; \; b = \frac{y_{max} + y_{min}}{2}, \tag{2}$$

where x_{max}, x_{min}, y_{max} and y_{min} are the maximum and minimum algebraic deviations of coordinates in a group of checked points taken independently of the fact to which actual point they refer.

The deviations of the coordinates of each point in the new system X' and Y' are determined from (3):

$$x' = x - a; \; y' = y - b. \tag{3}$$

In the case when the coordinate tolerances are equal, the component is passed by the positioning of its holes if the maximum (in its absolute value) ± deviation of its points along either of the axes is smaller than the tolerance, and the component is rejected if it is larger. If the tolerances in the coordinates of various points are not equal, the origin is displaced by the value of the tolerances and the measured deviation of the point coordinates.

The processing of the results by the parallel displacement of the origin is not sufficient, since it does not eliminate the error due to the turning of the measuring base.

In fact, all the points of a machined component are inevitably plotted with unknown errors. The setting points,

i.e., the points chosen for the measuring base, are also subject to errors. Owing to this circumstance it is possible to place various components in a different manner with respect to the actual system of coordinates and obtain in coordinate measurements different deviations for the same points of the component.

Let, for instance, two points A and B (Fig. 5) be chosen for the measuring base. It may be found after the boring of 3 holes with their centers at points A, B, and C that in fact the centers of two holes are close to the nominal value (points A and C), but the center of the third hole has a considerable displacement with respect to its nominal point B along both axes by the amount of Δx_B and Δy_B (Point B'). In this case after setting the component on the instrument for measuring the coordinates of points A and B', the component will be rotated through an angle $\Delta\varphi$ from its correct position. This is equivalent to saying that the component will not be measured in the coordinate system XY, but in a new system X'Y'. In order to eliminate this error it is necessary to rotate the system of coordinates and determine how the coordinates of the checked points are affected by this rotation.

The variations of the coordinates $\delta_1 x_p$ and $\delta_1 y_p$ of an arbitrary point P of the component (Fig. 6), when the coordinates X'Y' are rotated through a small angle $\Delta\varphi$ about their new origin, which coincides with point 1 of the component, are equal to

$$\begin{aligned}\delta_1 x_p &= x'_p \cos\Delta\varphi + y'_p \sin\Delta\varphi - x'_p = x'_p(\cos\Delta\varphi - 1) + \\ &\quad + y'_p \sin\Delta\varphi; \\ \delta_1 y_p &= y'_p \cos\Delta\varphi - x'_p \sin\Delta\varphi - y_p = \\ &= y'_p(\cos\Delta\varphi - 1) - x'_p \sin\Delta\varphi.\end{aligned} \tag{4}$$

For small angles of rotation $\cos\Delta\varphi \simeq 1$, and $\sin\Delta\varphi$ is directly proportional to $\Delta\varphi$. Reverting to the coordinate system XY and assuming that it is not the system X'Y' that is rotating but the component, we obtain an expression for the coordinate variations δx_p and δy_p of point P:

$$\begin{aligned}\delta x_p &= -(y_p - y_l)\cdot 0.00029\cdot\Delta\varphi\cdot 1000 = -0.29\,(y_p - y_l)\,\Delta\varphi, \\ \delta y_p &= +(x_p - x_l)\cdot 0.00029\cdot\Delta\varphi\cdot 1000 = 0.29\,(x_p - x_l)\,\Delta\varphi.\end{aligned} \tag{5}$$

where δx_p and δy_p are expressed in microns, x_p, y_p, x_l and y_l in millimeters, and $\Delta\varphi$ in minutes. The values of δx_p and δy_p are calculated [see (6)] from the nominal coordinate points, which is permissible, since these calculations produce a second order error which can be neglected.

$$\delta x_p = -0.29\,(y_{p_0} - y_{l_0})\,\Delta\varphi = k_1\,\Delta\varphi,\quad \delta y_p = 0.29\,(x_{p_0} - x_{l_0})\,\Delta\varphi = k_2\,\Delta\varphi. \tag{6}$$

Such calculations are made for all the points checked by their coordinate tolerances for rotation angle $\Delta\varphi$ values of 1' or 30" and an appropriate table is compiled. By means of this table it is possible to determine, from (7) and the coordinate deviations of the points measured at one setting of the component, the new coordinates of points for different settings of the component, and to arrive at correct conclusions whether to pass or reject the component.

$$x''_p = x'_p + \delta x_p,\quad y''_p = y'_p + \delta y_p. \tag{7}$$

It is recommended that in symmetrical components (for instance, a round or rectangular plate) which have a hole in the middle, the center of that hole be taken as the point of rotation. In asymmetrical components, such as cocks of watches, the center of the hole which serves to fix it to another detail should be taken as their center of rotation.

TABLE 1

Rotation angle $\Delta\varphi$	Variations in the coordinates of points, μ							
	III		IV		V		S_5	
	δx	δy	δx	δy	δx	δy	δx	δy
-1'	1.5	-1.7	1.9	-0.8	2.8	-0.6	3.8	0
0	0	0	0	0	0	0	0	0
+1'	-1.5	1.7	-1.9	0.8	-2.8	0.6	-3.8	0
+2'	-3.0	3.4	-3.7	1.7	-5.6	1.3	-7.6	0.1

Fig. 7

As an example let us calculate a table for the values of δx and δy for the centers of holes III, IV, V and S_5 of a gearing cock in the "Pobeda" watch, when it is rotated about the center of its base hole S_4 (Fig. 7).

The basic data for calculations consist of the nominal coordinates of the points. The base hole S_4 has nominal coordinates of $x_{S_4} = 109.360$, and $y_{S_4} = 106.500$.

Let us now calculate coefficients k_1 and k_2 for point III, which has nominal coordinates of $x_{III} = 103.435$ and $y_{III} = 101.394$:

$$k_1 = -0.29\,(y_{III} - y_{S_4}) = +1.48\ \mu;$$
$$k_2 = 0.29\,(x_{III} - x_{S_4}) = -1.72\ \mu.$$

Since variations in the point coordinates are directly proportional to the rotation angle $\Delta\varphi$, and for $\Delta\varphi = 1'$ they are $\delta x_{III} = 1.5\ \mu$ and $\delta y_{III} = -1.7\ \mu$, for an angle of $\Delta\varphi = 2'$ the variations will be twice as large, etc. For negative angles the coordinate variations will be of the same numerical value but with the reversed sign.

The values for δx and δy were calculated in a similar manner for the remaining points (see Table 1, shown in an abridged form).

TABLE 2

Number of the point	Processing of measurement results by means of a complete transformation of the coordinate system							
	Maximum deviations of points, μ		Measured deviations of point coordinates, μ		Processed deviations of point coordinates, μ			
					By displacing the origin of coordinates		By displacing the origin and rotating the coordinate system about point S_4	
	Along axis X	Along axis Y	Δx	Δy	$\Delta x'$	$\Delta y'$	$\Delta x''$	$\Delta y''$
1	2		3		4		5	
III	±6	±6	+12	-12	+8	-8	+5	-4.6
IV	± 5	±5	+12	-3	+8	+1	+4.3	+2.7
V	±5	±5	+13	-10	+9	-6	+3.4	-4.7
S_4	±4	±4	0	0	-4	+4	+4	+4
S_5	±4	±4	+15	-8	+11	-4	+3.4	-3.9

Next we determine whether to pass or scrap the component by the maximum deviations of the coordinates shown in Column 2 of Table 2.

Column 3 of Table 2 shows the measured deviations of the point coordinates.

Let us apply the displacement of the origin: $a = -4\mu$ and $b = +4\mu$ (Column 4). We find that for the given setting of the components the coordinate deviations $\Delta x'$ and $\Delta y'$ for the majority of the points exceed the tolerances. These calculations can be easily performed mentally. Let us now examine the component's angle of rotation $\Delta\varphi = +2'$. Column 5 shows the point deviations $\delta x''$ and $\delta y''$ for a new setting of the component which differs from the previous setting by an angle of $\Delta\varphi = +2'$. They are calculated in the following manner:

For point III:

$$\Delta x''_{III} = \Delta x'_{III} + \delta x_{III} = +8 - 3 = +5\ \mu;$$
$$\Delta y''_{III} = \Delta y'_{III} + \delta y_{III} = -8 + 3.4 = -4.6\ \mu.$$

The values of δx_{III} and δy_{III} for $\Delta\varphi = 2'$ are taken from Table 1.

The new deviations of the remaining points are calculated in a similar manner. The deviations of the points of the new setting of the component do not exceed tolerance. Hence, the component can be passed with respect to the position of these points.

The processing of measurement results by the above method of complete transformation of the coordinate systems, which was developed and introduced by the author at the Chistopol' watch-making plant, raises considerably the accuracy of checking the position of holes by the tolerances for the positions of their centers.

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INSTRUMENT FOR SELECTIVE SORTING OF SLIDING PAIR COMPONENTS

Ya. V. Aksenov, V. B. Kogan, A. P. Kurochkin, and A. B. Lind

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The existing methods of checking conjugate diameters of sliding pairs are nonproductive and do not provide the required accuracy, as a result of which the checked components have to be lapped to each other. This operation alters the dimensions and the geometrical shape of the details, thus leading to a reduced life of fuel pumps and a lowering of their hydraulic tightness.

In this connection the Interchangeability Bureau has developed a pneumatic instrument for an efficient and accurate checking of sliding pair details.

A low-pressure instrument mass-produced by the "Kalibr" plant was used for testing purposes. However, this instrument is designed for simultaneous checking of only two parameters and does not provide differential measurements, and lacks luminous signalling. The first difficulty was overcome by fitting the mass-produced instrument with a new distributor and additional manometric tubes. The second difficulty was overcome by means of a photoelectric transducer, whose schematic is shown in Fig. 1. The instrument is filled with distilled water tinted with a black dye-nigrosine (GOST 4014-48), a tubular lamp 1 is placed on one side of the measuring tube, and on the other, diaphragm 2 and a photoelectric transducer which consists of an ebonite plate with n photovaristors type FSK-5 mounted on it (their number depends on the quantity of checked parameters). Each photovaristor from 1 to n/2 is paired with photovaristors from (n/2)+ 1 to n respectively, forming two arms of unbalanced bridges. The other two arms of the bridges consist of resistors R. The diagonal of each bridge is connected to a polarized relay type RP-5.

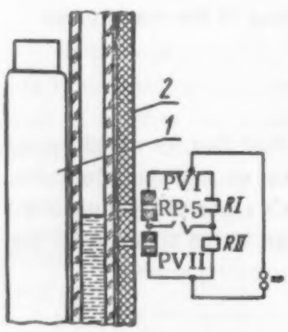

Fig. 1.

The bridge operates in the following manner. If both photovaristors are shaded, the current in the bridge diagonal is very small and the polarized relay armature remain in the neutral position. If the liquid column drops to such an extent that photovaristor PVI becomes illuminated, but photovaristor PVII remains shaded, the current in the bridge diagonal will increase and the polarized relay will operate, closing its right-hand-side contact.

When the liquid column drops still lower and photovaristor PVII becomes illuminated, the bridge diagonal will pass a current in the opposite direction (approximately twice as large in magnitude), and the polarized relay armature will close on its left-hand-side contact.

Despite the fact that such an operation of the polarized relay requires a considerable imbalance in the bridge, the stability of its operation is easily and reliably attained by selecting the required position of the resistor slider and by stabilizing the dc supply source. The closing by relay RP-5 of its left or right-hand-side contacts operates appropriate RKN type relays whose contacts light the required signal lamps.

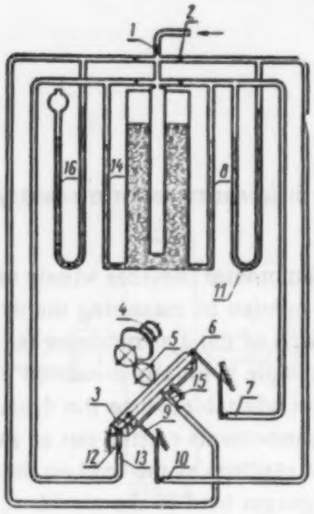

Fig. 2.

The use of photovaristors type FSK-5 with small overall dimensions (1 × 1 × 6 mm) provides a high sensitivity for the transducer. Thus, if the photovaristors are spaced by 1.8 mm in an instrument with a gain of 5400, its sensitivity will equal 0.3 μ. By means of such a photoelectric transducer it is possible to check the limits of articles, sort them according to several parameters, and check the regularity of their geometrical shape (tapering, ovality, flats and linearity of the generating line).

Figure 2 shows the schematic of an instrument for checking the following plunger parameters: its diameter with a sorting out of 25 groups in steps of 0.001 mm; its ovality (tolerance of 0.001 mm), its tapering (tolerance of 0.0015 mm), flats on its surface (tolerance of 0.001 mm) and linearity of its generating line (tolerance of 0.001 mm).

The compressed air passes from the mains through a filter and a reduction valve to throttle 1 and the water stabilizer of the instrument. The pressure-stabilized air is fed to nozzle 2 and then to the measuring nozzles and indicating tubes (liquid manometers). The contact levers of the measuring gear are suspended by flat springs. The measured component is placed on a 90° angular bearing 3, and is driven during measurements by motor 4 type RD-83 through a rubber roller 5 at a speed of 20 rpm. The knife-type contact lever 6, whose edge is parallel to the left-hand face of angle bearing 3, checks the plunger diameter. The diameter size of a given cross section detail controls the gap of the measuring nozzle 7 and hence the level in the corresponding indicating tube 8. Since the plungers are sorted according to their maximum diameters, the photoelectric transducer records the highest level of the liquid in tube 8 and lights the lamp corresponding to this group. The photoelectric transducer mounted on tube 8 also checks the plunger ovality as the maximum difference in diameters observed in the given cross section over a complete revolution.

The tapering of the plunger is determined as the difference of diameters in the two extreme positions. For this purpose contact lever 9 and nozzle 10 are used for checking in a similar manner the plunger diameter at the other end. The differences in the level of the liquid in manometer 11, which determine the tapering of the plunger, are recorded by means of a photoelectric transducer.

Contact lever 12 and nozzle 13 check the flats on the plunger (its tendency to a pentahedral shape), whose value corresponds to the variation of the level of the liquid in the indicating tube 14 over one rotation of the plunger, and is registered by an appropriate transducer.

The linearity of the plunger's generating line is checked by contact lever 15, whose edge is parallel to the right-hand face of the angle bearing. The deviations from linearity are registered on the indicating tube 16 and the corresponding transducer.

In the event of the tolerances not being met in any of the plunger parameters a reject alarm signal is given and the lamp corresponding to the diameter group does not light.

The setting and removing of the component in the measuring position, and its placing in the crate with a lighted lamp which corresponds to the diameter of the plunger, are done by the operator.

The device for checking the plunger pair, i. e., its cylinder, is basically the same as the above. The instrument checks the same parameters with the same tolerances, with the exception of the flats. The method of measurements is contactless, by means of a pneumatic plug.

The instruments are calibrated by means of samples. The operating cycle of each instrument is 6 sec.

Tests of the instrument's experimental model have shown that it provides an objective control of the plunger sliding pair parameters for a selective assembly, with an error of $\pm 0.3\,\mu$.

MECHANICAL MEASUREMENTS

DYNAMIC CALIBRATION METHOD FOR DYNAMOMETERS

A. S. Bol'shikh and L. G. Étkin

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Machines for testing fatigue in materials are often fitted with dynamometers of different types for measuring variable loads applied to the sample during testing.

Several methods are known for an indirect evaluation of dynamic errors in dynamometer devices widely used with fatigue-measuring machines [1, 2, 3]. One of the most commonly used methods consists in measuring the dynamic tension in the dynamometer and the sample simultaneously. The dynamic tension of the dynamometer is measured by means of the equipment comprised in the machine. The tension in the sample is usually measured by wire strain gauges. The replacement of the component by an "equivalent" detail is not advisable, since the dynamic characteristic of the machine's oscillatory system may thus be changed. From the measurements carried out at varying loads a graph is plotted that shows the relation of the machine's measuring instrument readings to the load on the sample. In view of the fact that such measurements are made by means of the strain gauges used in the machine, the error of such methods of evaluating the dynamic accuracy of the equipment is large.

Below we examine a method for dynamic calibration based on loading the tested dynamometers by inertial efforts whose value is determined by means of indirect measurements.

The schematic diagram of this equipment is shown in Fig. 1.

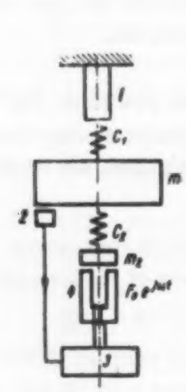

Fig. 1

The tested dynamometer 1 is rigidly fixed to the massive bedplate of the device. A component with stiffness C_1 is fastened in the dynamometer grip. The other end of this component is connected to a large mass m. It is advisable to have a set of such components, in order to be able to select appropriate stiffnesses and masses for making up oscillatory systems in such a manner that their partial frequencies remain in the range of the tested dynamometer operating frequencies. Moreover, it is desirable to have at least three check frequencies, at the beginning, in the middle and at the end of the range.

Mass m_1 is connected to stiffness C_2 and mass m_2. Mass m_2 must be considerably smaller than m_1, and the partial frequency of the system consisting of C_2 and m_2 must be considerably (5-10 times) greater than that of the system consisting of C_1 and m_1. The whole equipment is excited to vibrate in the first mode of the system's oscillations. This can be achieved by various means, for instance, the oscillations of mass m_1 produce in the electromagnetic transducer 2 a signal whose frequency is equal to that of the system's oscillations. This signal is amplified by power amplifier 3 and fed to the input of the vibration exciter of an electromagnetic (or an electrodynamic) type 4, which provides an effort varying according to the harmonic law at a frequency equal to that of the system's oscillations. Since the system's Q-factor is large ($2\cdot10^2 - 2\cdot10^3$) [6], the effort $F_0e^{j\omega t}$ is extremely small as compared with the effort which arises in the component with stiffness C_1. The high Q-factor of the mechanical oscillating system provides a small harmonic content in the oscillations of mass m_1.

Let us now examine the oscillatory movement of the described double-mass system.

The movement equations are of the form

$$
\begin{aligned}
&m_1\frac{d^2X_1}{dt^2}+k_1\frac{dX_1}{dt}+C_1X_1+C_2(X_1-X_2)=0;\\
&m_2\frac{d^2X_2}{dt^2}+k_2\frac{dX_2}{dt}+C_2(X_2-X_1)=F_0e^{j\omega t}.
\end{aligned}
\qquad (1)
$$

where X_1 and X_2 are the linear displacements of masses m_1 and m_2 respectively, $k_1 \cdot dX_1/dt$ and $k_2 \cdot dX_2/dt$ are the forces of resistance applied to masses m_1 and m_2, F_0 is the amplitude of the exciting force, ω is the lowest natural frequency of the double-mass system.

Effort N_2 arising in element C_2 is equal to

$$N_2 = C_2 (X_2 - X_1) = \frac{F_0 C_2 [(-m_1\omega^2 + C_1 + C_2 + j\omega k) - C_2]}{(-m_1\omega^2 + C_1 + C_2 + j\omega k_1)(-m_2\omega^2 + C_2 + j\omega k) - C_1 C_2}, \tag{2}$$

or

$$N_2 = C_2 (X_2 - X_1) = A + jB. \tag{3}$$

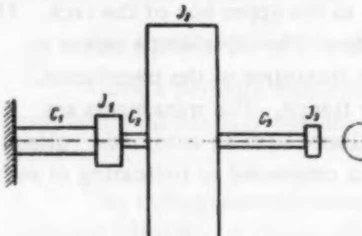

Fig. 2

This expression in a vectorial form represents the sum of two vectors, one of which is in the same phase as the displacements X_1 and X_2, and the other is shifted with respect to them by 90°.

If the system vibrates in resonance, effort F_0 differs in phase from displacement X_2 by 90° and the work is spent on supplying the lost energy in the system.

The inertial component of effort N_2 due to the movement of the mass is in phase with displacements X_1 and X_2, and therefore does not produce any work.

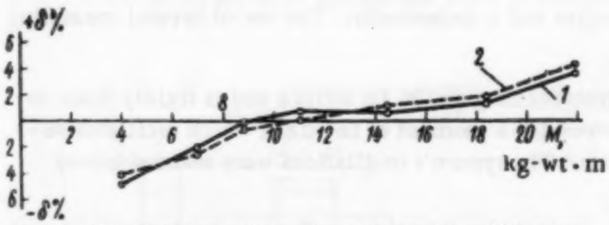

Fig. 3. 1) Dynamic calibration; 2) static calibration.

Effort N_2 transmitted to the system consisting of m_1 and C_1 is out of phase with respect to displacement X_1 by the angle of $\varphi = \sin^{-1}(F_0/N_2)$.

By means of the same reasoning we can find that effort N_1 which arises in element C_1 also consists of two components which differ in phase with respect to displacement X_1 of mass m_1 by the angle of $\varphi_1 = \sin^{-1}(F_0/N_1)$.

In the experiment the lowest oscillating frequency of the system is measured, as well as amplitude of oscillations of mass m_1, effort N_2 and the phase difference between efforts N_1 and N_2.

The frequency of oscillations can be easily measured with an error of 0.1%. Since the frequency of stable oscillations is measured, the simplest method consists in counting the number of pulses per unit of time from the sinusoidal voltage.

The oscillation amplitude is best measured by one of the optical methods, selecting the most suitable according to the value of the oscillations.

Effort N_2 can be measured with an error of 5-6% by means of a wire strain gauge glued to the component with stiffness C_2. Since this measurement is made for determining the calibration correction, the above error is permissible.

The phase difference between N_1 and N_2 can be measured by means of a loop oscilloscope or a phase meter.

The masses m_1 and m_2 or, in the case of torsion and bending oscillations, the corresponding moments of inertia, should be accurately determined before starting the experiment.

The effort applied to the dynamometer under test is

$$P = m_1 a_1 \omega_1^2 \sin \omega_1 t, \tag{4}$$

where a_1 is the oscillation amplitude of mass m_1, ω_1 is the natural frequency of the system for its first mode of oscillations*.

*In (4) $a_1\omega_1^2 \sin \omega_1 t$ represents the vibrational acceleration of mass m_1. If an accelerometer transducer is fixed on mass m_1 and, in determining effort P, the vibrational acceleration is measured directly, it becomes no longer necessary to measure amplitude a_1 and frequency ω_1 – Editor.

Effort N_2 and angle φ between N_1 and N_2 are measured in order to ascertain that

$$N_2 \sin\varphi \ll P \quad (5)$$

or to make the required correction in calibrating.

Condition (5) is usually easily met, since the Q-factor of the system is large. Hence, the unaccounted for forces used in overcoming the force of resistance are small, thus making the error in calibrating the dynamometer by this method also small.

The above method of calibrating dynamometers designed for measuring axial efforts can also be extended to torsion and bending oscillations.

Figure 2 shows a schematic of an oscillatory system during dynamic calibrations. Dynamometer C_1, whose elastic body consists of a 30KhGSA (HRC 36-38) steel tube, is firmly fixed by its flanges to the upper box of the rack. The same box carries on a bracket and a flange a microscope with an eyepiece micrometer. The microscope serves to measure the oscillation amplitude of the dynamometer component which carries the armatures of the transducers. The transducers themselves are fixed on rigid brackets mounted on the dynamometer flange. The transducers are connected to the electrical part of the dynamometer, which consists of a crystal oscillator with an automatic voltage control, a phase-sensitive detector and calibrated potential dividers whose outputs are connected to indicating or recording instruments [4,5].

The dynamometer grip, which is intended for holding the tested samples, is fixed to rod C_2 fitted with wire strain gauges glued to it at an angle of 45° to its axis (this rod provides stiffness C_1 in Fig. 1). A disc with a moment of inertia J_2 is firmly fixed to the other end of this rod. The moment of inertia of this disc is determined accurately by the method of suspending it by threads; in our experiment it amounted to 0.036 kg-wt·m/sec^2. The oscillation amplitude of the disc is measured by means of a set of measuring wedges and a cathometer. The use of several measuring wedges greatly raises the accuracy of measurement.

Another rod C_3 (which provides stiffness C_2 in Fig. 1) has transducers glued to its surface and is rigidly fixed to the bottom of the disc. The end of this rod is fixed to an armature with a moment of inertia J_3 which oscillates between the poles of a differential electromagnetic vibration exciter. The system's oscillations were maintained by means of amplifier TU-50.

The wire strain gauges glued to the rods were connected to the inputs of two channels of an eight-channel strain-gauge equipment 8AN4, and served to determine the phase difference between the torques operating in the rods.

Figure 3 shows the graph of deviations in the readings of a dynamometer indicating instrument taken for dynamic and static calibrations.

It will be seen from this graph that the differences between the static and dynamic calibrations are very small. The dynamically calibrated dynamometer was designed to work at frequencies up to 500 cps. It was calibrated at a frequency of 25 cps. An analysis has shown that with such a frequency ratio the error will be extremely small, and in practice the readings of the dynamometer in a static or dynamic condition should coincide. The above circumstance is confirmed by the cited graph. The existing nonlinearity is due to the tested dynamometer's registering device.

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ACCELEROMETRIC INSTRUMENTS

N. I. Kotel'nikov

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Instruments for measuring angular accelerations. In a number of instances it is necessary to measure angular acceleration in addition to angular velocity.

Figure 1 shows a schematic for measuring angular acceleration in the range of 0-400 and 0-2000 deg/sec^2 (the reduction amounts to 800 : 1).

The transducer consists of a tachogenerator type TMG-30P which also serves for measuring angular velocity.

Since a periodic zero setting is possible during measurements, the measuring device is made in the form of an asymmetrical parallel-balanced cathode-follower which uses tube 6N3P and provides such an adjustment by means of resistor R_4.

The tachogenerator voltage is differentiated by means of network C_1R_1 and fed to the control grid of the balanced cathode follower. A center zero instrument M-24 (grade 1.0) is used as an indicating device. Capacitor C_2 serves to reduce the effect of the tachogenerator pulsations on the instrument readings.

Resistor R_3 reduces the nonlinearity of the stage. The instrument scale is linear with an error up to 2%.

Switch S_1 serves to change the scale of measurements. The referred error of the instrument does not exceed 3%.

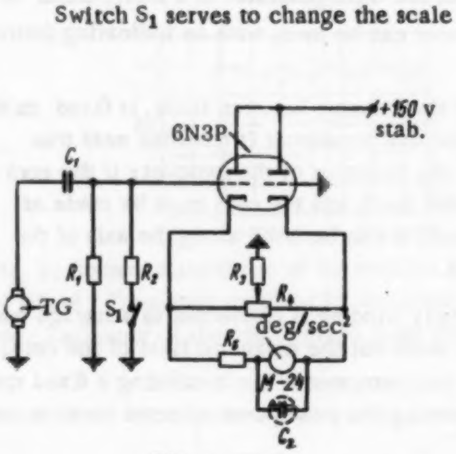

Fig. 1.

Instrument for measuring the speed of linear displacements in a reversible system. The measurement of the speed of linear displacements in a reversible system is beset with a number of difficulties which arise in differentiating, amplifying and rectifying slowly changing voltages.

Figure 2 shows the circuit of an instrument which by means of a parallel-balanced amplifier (its drift is random and sufficiently small), correcting networks and a sensitive measuring instrument has to a great extent eliminated a number of difficulties.

A wire potentiometer R_2 520 mm long is used as a transducer of linear displacement.

The voltage supplied by the transducer is differentiated by the C_2R_4 network and fed to the grid of the left-hand triode of the amplifier.

The voltage amplified by the left-hand-side triode is rectified and fed through a correction network consisting of R_7, C_4, C_3 and R_6 to the grid of the right-hand-side triode. The correction network with capacitors C_5 and C_6 provides a uniform scale for the instrument (the same purpose is served by the large multiplier resistor R_{11}) and in addition maintains during reversing the voltage across capacitor C_6 almost constant.

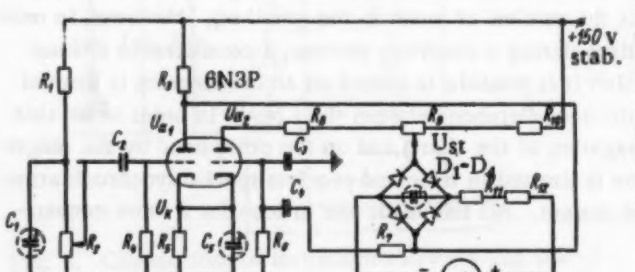

Fig. 2.

Resistor R_9 serves to set the zero. Diodes D_1-D_4 are of the D7Zh type. Diodes should be selected for maximum reverse resistances.

This circuit provides measurements of linear displacement speeds in the range of 0-50 cm/sec. By changing the value of potentiometer R_2 this range can be extended.

The dip in the pointer of instrument M-24 (grade 1.0) does not exceed 2% with reversals lasting up to 0.5 sec.

The referred error of the instrument does not exceed 2%.

The above instrument can be used in reversing systems for measuring the speed of linear displacements, for instance, in planing machines, various rocking devices, looms, etc. Up to the present the speed of linear displacements has either not been measured at all or measured indirectly. The instrument provides direct measurements of linear displacements, the readings being taken from a pointer instrument with a uniform scale.

This instrument considerably facilitates the work of machine setters and raises the productivity of their labor.

REGISTERING TACHOMETER

Yu. A. Zholkov, G. S. Okun', and B. V. Plaksin

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The most widely used instrument for measuring the speed of rotation (with or without recording it) is a tachogenerator. However, there are instances in which it is impossible to connect the tachogenerator to a shaft, either directly or by means of any transmission. In such a case an induction transducer can be used, with an indicating instrument consisting of a frequency meter, for instance, of the ICh-6 type.

A sprocket (Fig. 1) with the width of its teeth measuring about 0.7 of the distance between them, is fixed to the shaft whose speed of rotation it is required to measure. The core of the induction transducer is mounted near this sprocket in the plane of its rotations. The diameter of this core must equal the thickness of the sprocket; if the core exceeds that thickness its end must be tapered. The gap between the sprocket teeth and the core must be made as small as possible. If it is impossible to place the coil to one side of the shaft, it can be fixed along the axis of the shaft and projections cut out in the butt of the shaft (or in a separate detail).

The induction transducer has a supply and a sensing winding. The supply winding is connected to a storage battery, and the sensing to the input of a frequency meter. When the sprocket teeth cut the magnetic field of the core, pulses are produced in the sensing winding whose frequency is measured by the instrument. For measuring a fixed speed of rotation, such a method is convenient. The error can be reduced by recording the pulses over selected portions on an oscillograph and counting their number per unit of time.

It is very difficult, however, to use this method for measuring variable speeds of rotation. The oscillogram records a large quantity of sawtooth pulses whose number per unit of time, marked on the oscillogram by a time pip, determines the required speed of rotation (taking into account the number of teeth on the sprocket). However, in order to obtain the relationship between the speed of rotation and time during a transitory process, a considerable amount of work is required. Moreover, the duration of the process which it is possible to record on an oscillogram is limited on the one hand by the requirement to record pulses with a sufficient distance between their peaks in order to be able to process the oscillogram (this determines the speed of propagation of the chart), and on the other hand by the length of the chart. Finally, such a recording of the speed of rotation is limited in time and requires special synchronization in order to be able to connect the oscillograph at the required instant. All this leads one to look for a more convenient method of recording.

Below we describe a method of recording the speed of rotation by measuring the mean value of the rectified current at the output of the frequency meter type ICh-6. It is known that this instrument measures the mean value of the rectified current in a circuit with a capacitor which is charged at the rate of the measured frequency. For an indicating instrument the ICh-6 set uses a sensitive microammeter whose pointer's angle of deflection is directly proportional to the number of discharges and charges per second. Figure 2 shows an approximate form of the voltage at the output of the ICh-6 set, which is fed to the recording instrument type ÉPP-09. The place where leads 9 and 10 are connected to the circuit of the ICh-6 set is shown in Fig. 3. The connection of electronic potentiometer ÉPP-09

in parallel with the microammeter introduces a substantial error in the readings of the latter, making it impossible to use its readings. This in any case is no longer necessary, since ÉPP-09 becomes an indicating and recording instrument.

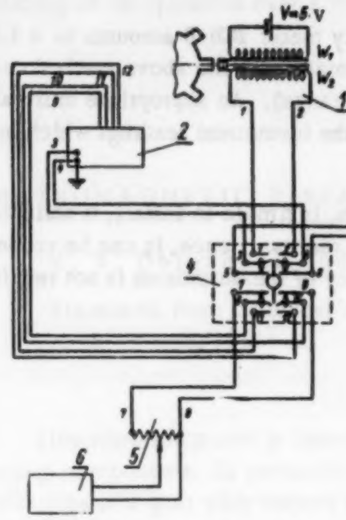

Fig. 1. Circuit for measuring and recording the speed of rotations. 1) Induction transducer; 2) frequency meter ICh-6; 3) oscillator ZG-10; 4) switch PD-6; 5) potential divider (R = 0-470 ohm); 6) potentiometer ÉPP-09.

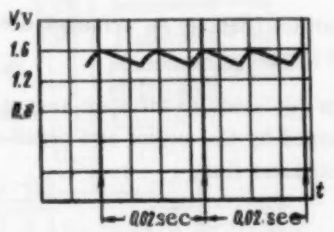

Fig. 2. Approximate shape of the voltage at the output of ICh-6. Frequency supplied by oscillator ZG-10, 100 cps; frequency of the oscillograph time pips, 50 cps.

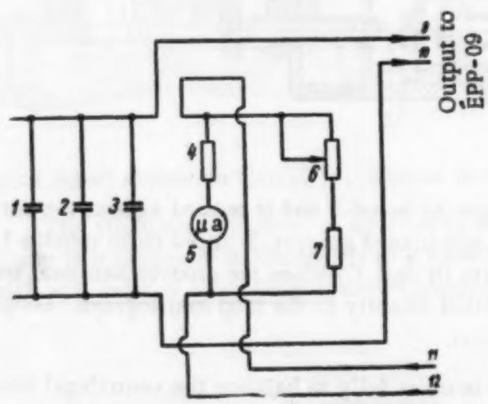

Fig. 3. Connections of instruments ÉPP-09 and ICh-6. 1, 2, 3) Capacitors (71, 72, 73); 4) wire-wound resistor (74); 5) microammeter M-24 (75); 6) wire-wound potentiometer (76); 7) wire-wound resistor (77). Figures in parentheses denote numbers on the schematic attached to technical description of meter ICh-6.

The instruments' layout for calibrating the speed of rotation on potentiometer ÉPP-09 and for checking instrument ICh-6 (Fig. 1) shows switch PD-6 in its middle position, used for calibrating set ÉPP-09. The output of oscillator ZG-10 is connected by means of leads 5 and 6 to the switch, and from it by leads 3 and 4 to the input of ICh-6. The output of ICh-6 is connected by means of leads 9 and 10 to the switch, and from it by leads 7 and 8 to ÉPP-09. In the top position of switch PD-6 contacts 9 and 10 are connected to 11 and 12, and contacts 3 and 4 remain connected to contacts 5 and 6. This position can be used for checking the readings of the microammeter of set ICh-6 (which is only in circuit when contacts 9 and 10 are connected to contacts 11 and 12) at a frequency of 10 kc supplied by oscillator ZG-10.

In the operating conditions for measuring and recording the speed of rotation switch PD-6 is in its lower position. In this position contacts 1 and 2 are connected to 3 and 4, and contacts 7 and 8 to contacts 9 and 10. The output of the induction coil is connected to the input of frequency meter ICh-6 and then to the recording instrument.

In the transducer used by us the sensing winding had 3200 turns of 0.25 mm^2 ÉP (enamelled) wire, the supply winding had 2500 turns of 0.35 mm^2 ÉP wire. The supply voltage (V = 5v) was obtained from cadmium-nickel alkali batteries type 5 NKN-45. The coil core diameter was 14 mm. The gap between the core and sprocket was 1-2 mm. The wires between the induction coil and frequency meter ICh-6 must be carefully screened to avoid the effect of any extraneous fields. The voltage at the output of ICh-6 is of the order of several volts. For recording purposes potentiometer ÉPP-09 was used, whose XA scale is calibrated for measuring temperatures of 0-800°C, which correspond to a maximum input voltage of 33.32 mv. The minimum number of changes were introduced into the ÉPP-09 circuit; thus the unit for compensating the thermocouple cold junction was disconnected and an additional resistance was introduced. If the top speed of rotation is not known, or if the instrument may have a varied application, it is advisable to add a variable resistor. Since it is necessary to attenuate the voltage and provide a great accuracy of regulation, it is advisable to use a variable resistor in the form of a potential divider (Fig. 1).

In our experiments of recording the speed of rotation the ÉPP-09 chart was propelled at the rate of 4800 mm/hr. All the 12 channels of the instrument were connected for recording one reading, thus making it possible

to record, with a high rate of chart propulsion and a recording cycle of 0.75 sec, very rapid variations in the speed of rotation.

If it is taken into consideration that the error of measurement of frequency meter ICh-6 amounts to ± 1.5%, and that of potentiometer ÉPP-09 to ± 0.5 %, the error of measuring the speed of rotation by the above method can be evaluated at ± 2% of the full scale reading (for a given setting of the ICh -6 range). An appropriate calibration of potentiometer ÉPP-09 provides a relation between the speed of rotation and the instrument readings which approaches linearity.

Conclusions. The above circuit consists of standard measuring instruments, is simple to install, is reliable and convenient in operation, and does not require additional processing of the test results. Hence, it can be recommended for the recording of constant and variable speeds of rotation when high accuracy of measurements is not required.

TURBOBLOWER SPEED OF ROTATION PULSE TRANSDUCER

V. N. Petrov, A. G. Savel'ev, and G. D. Silukov

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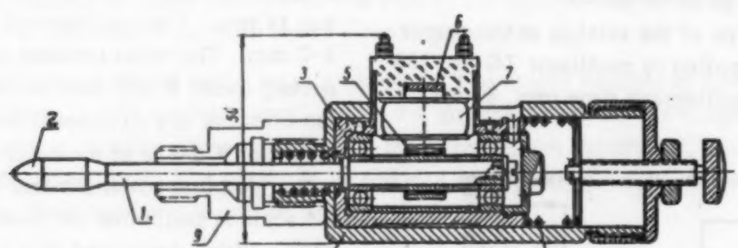
The testing of internal combustion engines equipped with turbochargers requires an accurate determination of the turbocompressor's (GTK) speed of rotation, which attains over 30,000 rpm.

When a multiloop oscillograph is used for recording test results it is advisable to measure accurately the speed of rotation by means of the pulse induction transducer (see figure) developed by the author and based on the well-known principle of inducing an emf in a closed circuit by varying the magnetic flux.



Transducer spindle 1 with its rubber tip 2 is mounted in a moving barrel 3 and is pressed against the butt of the turbo-charger shaft by spring 4. Sleeve 5, with its two symmetrically placed grooves, is fitted on to spindle 1 and rotates with it. Barrel 3 carries a telephone earpiece magnet 6 with its coil 7. When the grooves pass near the permanent magnet poles an emf is induced in the coil and transmitted directly to the loop oscillograph. Loops of various sensitivities can be selected to fit the speed of the shaft rotation.

Two symmetrically placed grooves in sleeve 5 are required in order fully to balance the centrifugal forces in the rotating transducer. Each turn of the shaft produces two oscillations in the loop, equivalent to the number of grooves in sleeve 5. Spring 4 provides a smooth coupling of the transducer to an operating turbocharger, and a reliable contact between spindle 1 and the turbocharger shaft.

The transducer casing 8 is provided with a thread for fixing it to turbochargers. With an adapter fitting 9 the transducer can be used for measuring the speed of rotation of other mechanisms.

The lubrication of bearings and other friction components of the transducer is made through openings in barrel 3. Excess oil percolates from the barrel into the cavity of spring 4 through the holes in the barrel lid. When measuring a stable speed of rotation this transducer can be used in conjunction with commercially produced pulse counters.

Extended use of the transducer in thermotechnical testing of marine engines with turbochargers has shown the reliability of its operation over a wide range of rotational speeds.

ELECTROMAGNETIC VIBRATION METER

V. S. Dolgachev and V. A. Shchinyavskii

Translated from Izmeritel'naya Tekhnika, No. 3,
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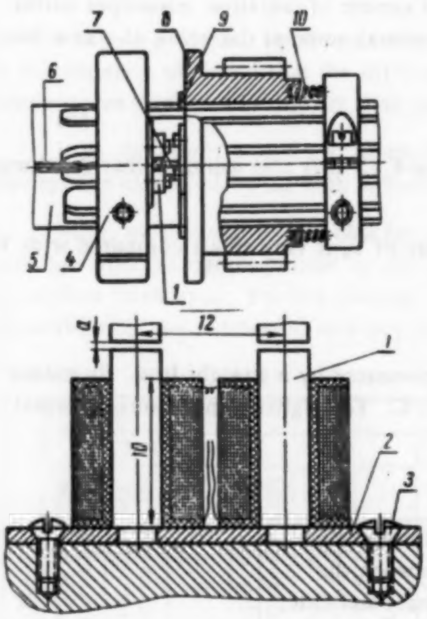
This vibration meter is intended for determining the magnitude of relative vibrations between two conjugate rotating components. In particular the vibration meter is suitable for determining reliably the existence of rotational oscillations of a gear with respect to the keyed shaft on which it is fitted.



The principle of operation of the meter is based on the changes in the relative position of the shaft and the gear within the limits of their free play which produce proportional variations in the magnetic flux and induce an emf in the transducer windings.

The principal part of the vibration transducer (figure) consists of electromagnet 1 fixed to flange 2. Thin wire (0.03-0.05 mm) is used for winding the coils, thus providing a large number of turns for a small size and raising the sensitivity of the transducer. By means of screws 3 the vibration transducer is fixed to split ring 4, which is mounted on shaft 5 by a coupling screw. The ends of the winding are taken along the key recess and groove 6 cut in the shaft, out of the reducer to the slip rings which are fixed on an ebonite cylinder tightly fitted on the end of the shaft, thus maintaining it in a fixed position.

The emf pulses are fed from the slip rings through brushes to the oscillograph loops.

Magnetic core 7 of a special shape is fixed to gear 8. The required gap a in the magnetic circuit is set by means of ring 4. The gap amounts to 0.3-0.5 mm.

The position of the gear is fixed by means of a check ring 9 and six (or three) expansion springs 10. Instead of the check ring 9 it is possible to use ordinary distance rings.

The vibration transducer in addition to detecting free play and backlash can also evaluate their size. For this purpose it is necessary to connect it to a known gap and plot a calibration curve of vibrations for the same parameters which will exist in testing the unknown coupling.

THERMOTECHNICAL MEASUREMENTS

RELATION OF THE THERMAL EMF TO TEMPERATURE IN TELESCOPES TYPE TERA-50/900-1800

E. S. Shpigel'man and L. M. Golub

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pp. 16-17, March, 1961

In an ideal case when the construction of a radiation pyrometer is simple and does not include any reflecting or refracting systems, the thermal emf for any wavelength is expressed as a function of temperature by the relation

$$e = a\,(T^4 - T_0^4)\,, \tag{1}$$

where T is the temperature of the source (black body), T_0 is the temperature of the receiver.

Since under actual conditions the total transmission through the optical system of radiation telescopes varies considerably with T, condition (1) is difficult to maintain, and according to several authors, the value of e as a function of T can be expressed with better approximation by

$$e = a\,(T^b - T_0^b), \tag{2}$$

where b is an index, constant for each instrument and varying between 3.5 and 4.5. It is also assumed that the dimensions of the source and its distance from the telescope remain constant.

For pyrometers with a lower measuring limit at or above 900°C the value of T_0^b is very small compared with T^b and in practice the following equation holds:

$$e = aT^b. \tag{3}$$

The calibration curve plotted from (3) in logarithmic coordinates is represented by a straight line. By means of it radiation pyrometers can be calibrated up to a temperature of 1300-1400°C. For higher temperatures thermal emfs are obtained by extrapolation of this straight line.

TABLE 1

Temperature, °C	Mean values of the measured thermal emfs, mv	Values of thermal emfs calculated by the least squares method from (4), mv	Discrepancies between the meas. and calc. values of thermal emfs, expressed in °C
1	2	3	4
900	2.34	2.32	-2
1000	3.82	3.83	+0.5
1100	5.95	5.99	+1
1200	8.91	8.91	0
1300	12.78	12.72	-1
1400	17.53	17.54	0
1500	23.52	23.51	0
1600	30.85	30.75	-1.4
1700	39.28	39.38	+1
1800	49.35	49.57	+2

The experience of the KhGIMIP (Khar'kov State Institute of Measures and Measuring Instruments) in calibrating radiation telescopes RP, Siemens ardometers and other instruments has shown that there is a considerable difference between the curve of the relation of log $\underline{e}$ and log T obtained experimentally and calculated from (3), and that this difference exceeds the calibration errors. In this connection grade 2 reference radiation telescopes are now calibrated directly from a "black body" radiator by means of a grade 1 reference pyrometer.

Our industry has now developed new types of radiation telescopes.

The authors of this article have set themselves the task of finding on the basis of experimental data an analytical expression which provides a sufficiently accurate representation of the relation between the thermal emf and temperature. This in turn will simplify the technique of calibrating grade 2 reference telescopes. For this purpose we investigated 10 radiation pyrometers type TERA-50 with a glass lens and a sighting index of 1/20.

The mean value of the measured thermal emfs for various temperatures is given in Table 1.

For the solution of our task we processed the data of column 2 in Table 1 by the least squares method expanding expression (3) into a series and substituting for the sake of convenience T°K by t°C. The analysis has shown that a mean calibration curve is most accurately expressed by a fourth-power equation:

$$e = at + bt^2 + ct^3 + dt^4. \tag{4}$$

Next we processed by means of (4) and the least squares method the calibration results of various telescopes. In this instance we found that the difference between the observed thermal emfs and those calculated from (4) by the least squares method was smaller than the calibration error of grade 2 telescopes.

It is, therefore, possible to assert that the relation between the thermal emf and temperature for telescopes of that type is expressed by (4) with sufficient accuracy.

Next, we decided to find how far the curves represented by (4) and plotted from four measured points would coincide with the curves plotted by the least squares method and with the experimental calibration curves obtained for various telescopes. For this purpose we selected a series of radiation telescopes type TERA-50 of the abovementioned version and calibrated at every 100°C in the range of 900-1800°C.

TABLE 2

	Telescopes																				
	№ 1			№ 2			№ 3			№ 4			№ 5			№ 6			№ 7		
t, °C	exptl. thermal emf values, mv	calc. thermal emf values, mv	Δt, °C	exptl. thermal emf values, mv	calc. thermal emf. values, mv	Δt, °C	exptl. thermal emf values, mv	calc. thermal emf. values, mv	Δt, °C	exptl. thermal emf. values, mv	calc. thermal emf values, mv	Δt, °C	exptl. thermal emf values, mv	calc. thermal emf values, mv	Δt, °C	exptl. thermal emf values, mv	calc. thermal emf values, mv	Δt, °C	exptl. thermal emf values, mv	calc. thermal emf values, mv	Δt, °C
900	2.23	2.29	−4	2.48	2.53	+5	2.46	2.45	0	2.26	2.28	+2	2.28	2.29	0	2.13	2.10	−3	2.28	2.24	−4
1000	3.82	3.82	0	4.01	4.01	0	4.07	4.07	0	3.69	3.69	0	3.78	3.78	0	3.68	3.68	0	3.80	3.80	0
1100	5.90	5.98	+2	6.30	6.24	−3	6.21	6.32	+4	5.70	5.71	0	5.90	5.92	+1	5.78	5.83	+2	5.90	5.97	+3
1200	8.88	8.88	0	9.34	9.31	0	9.34	9.34	0	8.48	8.48	0	8.89	8.89	0	8.66	8.66	0	8.86	8.86	0
1300	12.71	12.64	+1.5	13.48	13.41	+1.5	13.36	13.27	−2	12.16	12.13	−1	12.59	12.60	0	12.39	12.31	−2	12.71	12.60	−3
1400	17.36	17.36	0	18.54	18.54	0	18.26	18.26	0	16.80	16.80	0	17.36	17.36	0	16.91	16.91	0	17.31	17.31	0
1500	23.13	23.21	−1	24.64	24.81	+2	24.59	24.48	−1	22.61	22.64	+0.5	23.34	23.23	−2	22 71	22.65	−1	23.03	23.14	+1
1600	30.27	30.27	0	32.30	32.30	0	32.10	32.10	0	29.82	29.82	0	30.33	30.33	0	29.70	29.70	0	30.23	30.23	0
1700	38.75	38.70	0	41.30	41.06	−2	40.80	41.31	+5	38.02	38.49	+5	38.78	38.77	0	37.73	38.23	+6	38.35	38.79	+4
1800	49.01	48.64	−4	51.80	51.15	−6	51.35	52.30	+9	47.63	48.83	+10	48.68	48.68	0	47.53	48.68	+7	48.03	48.82	+7

The analysis of the possible conditions has shown that the best temperatures for the four points of curve (4) are 1000, 1200, 1400 and 1600°C (for the telescopes under investigation). From the values of the thermal emf at the above temperatures a system of simultaneous equations was solved for each telescope:

$$e_1 = a1000 + b1000^2 + c1000^3 + d1000^4, \qquad e_3 = a1400 + b1400^2 + c1400^3 + d1400^4,$$
$$e_2 = a1200 + b1200^2 + c1200^3 + d1200^4, \qquad e_4 = a1600 + b1600^2 + c1600^3 + d1600^4, \tag{5}$$

from which we obtained coefficients a_i, b_i, c_i and d_i for all the telescopes.

Table 2 shows the values of the thermal emfs obtained from experimental calibrations of seven telescopes by means of "black body" radiations and those calculated from coefficients a_i, b_i, c_i and d_i obtained by solving equation (5). The table also shows the values of Δt equal to the difference between the experimental and calculated values of the thermal emfs expressed in terms of °C.

From the above data it follows that the value of Δt is considerably smaller than 2σ, where σ is the quadratic mean calibration error of grade 2 radiation telescopes equal to ±2.5°C in the range of 900-1300°C, and to ±4°C in the range of 1400-1800°C. Only in the case of two telescopes does Δt exceed slightly the value of 2σ, at a temperature of 1800°C.

Conclusions. As the result of the above investigation we obtained a mean calibration curve and an analytical equation which represents with sufficient accuracy the relation between the thermal emf and temperature for the types of telescopes we investigated. This makes it possible to recommend a new, more efficient technique of calibrating grade 2 telescopes at four temperatures in the range of 900-1800°C.

CERTAIN ERRORS IN CHECKING COMMERCIAL RESISTANCE THERMOMETERS

B. K. Bragin

Translated from Izmeritel'naya Tekhnika, No. 3, pp. 17-19, March, 1961

The most important and difficult operation in checking commercial resistance thermometers consists in determining R_0 and R_{100}, the values of the sensing element resistance at 0 and 100°C. The permissible deviation of R_0 and of the ratio R_{100}/R_0 from their nominal values amounts to as little as 0.05-0.1%. Hence, even small errors in the calibration technique will lead to a false conclusion about the serviceability of the tested thermometer.

Experience has shown that the most widespread reason for an unjustifiable rejection of resistance thermometers, especially platinum thermometers, consists in the fact that owing to unsatisfactory heat exchange conditions equal temperatures are not provided for the thermometers' sensing elements and the ambient medium consisting of melting ice or water vapor. The errors thus incurred have a systematic nature, always making the value of R_0 higher than it should be, and the values of R_{100} and R_{100}/R_0 lower than they should be. Special measurements (see table) on two standard-type platinum thermometers with one and two sensing elements confirm the above assumption and provide an idea of the value of possible errors.

The immersion depth of the thermometer into a melting ice or steam thermostat is the basic factor determining the heat exchange condition. Experiments were, therefore, carried out at three immersion depths of 200, 250 and 300 mm, measured from the lowest point of the sensing element.

An insufficiently careful packing of the ice before measuring R_0 leads to the formation of a small (1-2 mm) air and water gap between the test tube thermometer guard and the mass of melting ice. In the calibration of commercial thermometers this circumstance is often overlooked. The figures in the table in parentheses show the measurements of R_0 obtained in the presence of this gap between the test tube and the surrounding mass of ice.

For the protection of the thermometers' sensing elements during testing three types of test tubes were used 350 mm long with an internal diameter of 12 mm and a wall thickness of 1 mm.

The measuring current in the thermometer in all the cases was maintained close to 5 ma. The resistance was measured with an error not exceeding 0.01% by means of a balancing device.

The analysis of the experimental data shown in the table makes it possible to arrive at certain important conclusions of practical interest.

1. The error in determining the resistance of the thermometer due to its insufficient immersion (200 mm) into a melting ice or steam thermostat amounts to 0.05% when glass guard tubes are used, and to 0.02% with aluminum tubes. The material of which the tubes are made and their shape do not appear to be important when the thermometer is immersed to a depth greater than 300 mm measured from the bottom of the sensing elements. This should be considered as the minimum immersion in testing the thermometers.

2. The relative errors in R_0 and R_{100} due to the insufficient immersion of thermometers are close in value but have opposite signs. Hence the value of the ratio R_{100}/R_0 depends to a marked extent on the depth of immersion.

3. The formation of a gap due to the melting of ice round the guard tube raised the measured value of R_0 by 0.1, 0.05 and 0.02% when the thermometer was immersed to 200, 250 and 300 mm respectively. Thus, the packing of the ice mass is absolutely necessary even with a thermometer immersion of 300 mm. This is all the more important because most resistance thermometers, for reasons of a technological nature, are produced with a small positive deviation of R_0 from the nominal value, which makes the probability of wrong rejection even greater.

Material and shape of tubes	Thermometer no.		Depth of immersion, mm								
			200			250			300		
			R_0	R_{100}	R_{100}/R_0	R_0	R_{100}	R_{100}/R_0	R_0	R_{100}	R_{100}/R_0
Glass, cylindrical	№ 1		46.04_4 (46.08_1)	63.90_3	1.387_8 (1.386_1)	46.03_3 (46.07_2)	63.92_3	1.388_8 (1.387_4)	46.02_0 (46.04_0)	63.92_4	1.388_8 (1.388_1)
	№ 2	I	46.03_4	63.80_8	1.386_1	46.01_8	63.81_9	1.386_8	46.01_0	63.83_0	1.387_3
		II	46.05_3	63.83_1	1.386_1	46.03_9	63.84_4	1.386_0	46.03_0	63.86_1	1.387_4
Aluminum, flattened at the lower end	№ 1		46.03_0 (46.05_8)	63.91_6	1.388_8 (1.387_8)	46.02_8 (46.03_8)	63.92_8	1.388_8 (1.388_0)	46.02_3 (46.03_0)	63.92_3	1.388_9 (1.388_7)
	№ 2	I	46.01_8	63.81_9	1.386_9	46.01_3	63.82_8	1.387_1	46.01_1	63.82_8	1.387_9
		II	46.03_3	63.84_8	1.386_8	46.03_0	63.85_3	1.387_8	46.03_0	63.85_0	1.387_1
Aluminum, factory jacket	№ 1		46.03_3	63.91_3	1.388_4	46.03_3	63.92_0	1.388_8	46.02_0	63.91_8	1.388_7
	№ 2	I	46.02_0	63.81_7	1.386_7	46.01_4	63.82_3	1.387_3	46.00_7	63.82_7	1.387_3
		II	46.03_3	63.84_7	1.386_9	46.02_8	63.85_3	1.387_8	46.02_4	63.86_4	1.387_8

The comparison of the data given above may lead to the conclusion of the desirability of using metallic guard tubes, in particular aluminum factory-produced jackets of 350 mm long thermometers (providing they are hermetically sealed). One should not, however, overlook the possibility of condensation and accumulation of moisture inside the jacket when it is immersed in an ice bath. This can lead to additional errors in measuring the resistance of the thermometer. It is, in practice, impossible to ascertain the existence or absence of moisture inside a metallic jacket. They should, therefore, be used with great care and cannot be recommended for extensive testing. In addition to the above, two more sources of errors should be mentioned.

In checking commercial resistance thermometers laboratory potentiometers are used, whose basic error lies in the limits of 0.015-0.030%. When the values of V_t and V_r (V_t and V_r are the voltage drops across the thermometer and the reference resistor respectively) are equal it is possible to neglect corrections to the readings of the potentiometer. However, in the majority of cases V_t and V_r differ from each other to such an extent that the corresponding readings of first potentiometer decade are not equal. Hence, the value of V_t/V_r may contain an error which cannot always be neglected. It is necessary to evaluate in advance from the potentiometer certificate of repair the possible error in determining V_t/V_r. If it amounts to more than 0.01%, the potentiometer readings should be appropriately corrected.

In order to determine the value of R_0 with an error not exceeding 0.01% it is necessary to measure the temperature in the vapor thermostat with an error not exceeding 0.01%. For this purpose reference mercury-in-glass grade 2 thermometers are used. It is absolutely essential to correct the readings of these thermometers, since their errors

may attain ±0.2°C. At the same time it is necessary to take into consideration the variation in the reference thermometer corrections due to their natural aging.

MINIATURE RESISTANCE THERMOMETERS FOR CHECKING THE OPERATION OF BEARINGS

V. V. Ipatov and I. Ya. Magin

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pp. 19-20, March, 1961

The operation of turbogenerator bearings is usually checked by the temperature of the overflow oil from the bearings.

However, this method does not provide a reliable check on the operation of the bearings, since even a considerable heating up of the oil in the lubricating layer raises the temperature of the overflow oil only slightly. Moreover, the latter temperature rises with a lag, thus making it impossible to guarantee a timely signal for the prevention of any damage. This is due to the fact that the amount of oil which flows through the lubricating layer is considerably smaller than that flowing through the bearing as a whole. Neither does this method ensure the checking of the temperature of each thrust bearing block.

A more reliable method of checking radial and axial bearings consists in measuring directly the temperature of the babbitt lining.

Tests have shown that the temperatures of bearing linings vary, depending on the construction of the bearing, the load, and the quantity and temperature of the cooling oil, in the range of 70-90°C, attaining in a danger condition the temperature of 110-140°C. Therefore, the equipment for checking the operation of bearings should be designed to measure temperatures up to 150°C.

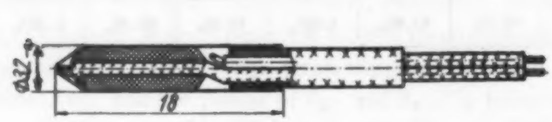

Fig. 1

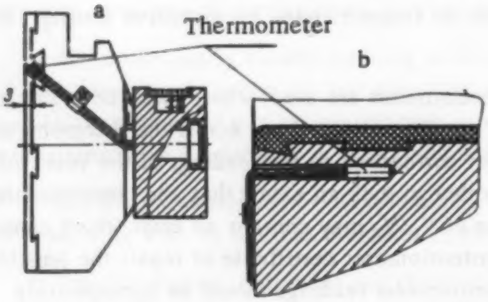

Fig. 2

For this purpose it is advisable to use resistance thermometers. However, the resistance thermometers mass-produced by our industry are rather large and cannot, therefore, be placed in thrust bearing blocks. In this connection the laboratory of the turbomotor plant has designed a miniature resistance thermometer 3.2 mm in diameter and 12 mm long in its effective part.

This resistance thermometer (Fig. 1) consists of copper wire (diameter 0.05 mm) type PÉL GOST 2773-51 wound in a bifilar manner directly over one of the thermometer leadout conductors and covered with a layer of bakelized paper. The thermometer resistance at 0°C is equal to 53 ohm; its calibration corresponds to that of 2a copper thermometers.

Experiments have shown that the stability and measurement error of these thermometers meet the GOST 6651-59 requirements for grade III thermometers.

Miniature thermometers have a very quick response, which is very important when they are used as protection devices in bearings.

The position of a miniature resistance thermometer in a thrust bearing block of turbine type VPT-25-4 and in the thrust bearing lining is shown in Fig. 2, a and b.

The resistance thermometers are placed in specially drilled holes (diameter 3.4 mm) and are glued with type BF-2 adhesive. The lead-out conductors of the resistance thermometers are taken through plug connections to a switch and a measuring instrument mounted on a separate board.

Electronic bridge ÉMV-11 is used as a measuring instrument. By means of this instrument and a switch it is possible to check periodically at a distance the temperature of thrust bearing blocks and lining, thus controlling the distribution of load between the blocks. An electronic bridge type ÉMDS-26 can also be used if its platinum thermometer calibration (11a) is replaced by a copper thermometer calibration (2a). When instrument ÉMDS-26 is used the current through the resistance thermometer becomes a little higher than with the ÉMV-11 instrument; however, tests have shown that this does not produce a noticeable rise in temperature due to the current. The application of miniature thermometers raises the reliability in checking the behavior of bearings under working conditions, and makes it possible to ascertain whether the thrust bearing blocks have been correctly fitted in assembly.

EXPERIENCE GAINED IN USING SEMICONDUCTOR RESISTANCE THERMOMETERS FOR REMOTE MEASUREMENTS OF TEMPERATURE

F. A. Tserikh

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 20, March, 1961

Semiconductor resistors (thermistors) can be used, providing they are stable, as resistance thermometers for remote measurements of temperature with high accuracy.

Thermistor thermometers were used by us for observing the stability of temperature at a depth of 15 m in a special well which was constructed a long time ago on instructions of D. I. Mendeleev and is now used as a thermostat for crystal oscillators.

For this purpose we selected out of several thermistors of different types, after their thermal aging and checking of their stability, two of the most stable thermistors (MMT-6 and MMT-4) and used them as resistance thermometers.

The above thermometers were calibrated in a water thermostat against high-precision reference mercury-in-glass thermometers. The resistance of the thermometers both during their calibration and subsequent temperature measurements in the well was measured with a laboratory type bridge MVL-47 equipped with a highly sensitive null-galvanometer GPZ-2 at a measuring current of 30 μa.

The stability of the thermometer readings was systematically checked by measuring their resistance at the triple point of water (+ 0.01°C).

From these measuring results the following conclusions can be drawn:

1. The stability of thermistor thermometers over the testing period was found to be completely satisfactory; the characteristic quadratic mean errors were ± 0.0022°C (MMT-4) and ±0.006°C (MMT-6).

2. Thermistor thermometers can be used after careful selection and checking of their stability for remote measurements of temperature with high accuracy. Both in these measurements and during calibration a constant measuring current through the thermometer must be provided.

ELECTRICAL MEASUREMENTS

STABILIZATION OF SMALL DIRECT CURRENTS BY MEANS OF STANDARD CELLS

Z. I. Zelikovskii

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pp. 21-24, March, 1961

Experimental investigations of the behavior of standard cells under load [1-4] have shown that they can be used for electrical measuring circuits with a current flowing through them. As an example of such an application we can cite the new method of comparing standard cells by means of a calibrated galvanometer [3], the use of standard cells as a source of a stable operating current in high-precision compensators, and as reference elements in compensating voltage stabilizers [5].

In the present work we examine the application of standard cells in a floating condition for long-term stabilization of small direct currents.

Schematic and computation of a stabilizer. The schematic of a stabilizer with a standard cell operating in a floating condition is shown in Fig. 1. The nominal (initial) values of the emf and resistances in the circuit are selected in such a manner that current I_2, which flows through the standard cell E_2, is equal to zero. Hence, the source of the stabilized current I, which flows through the load resistor R, consists of E_1.

$$I=\frac{E_1}{R_1+R}=\frac{E_2}{R}, \tag{1}$$

where R_1 is the ballast resistance, which includes the resistance of source E_1.

If after a lapse of a certain time or under the effect of certain factors the parameters of the circuit deviate from their initial values, current I will change and current I_2 will appear. It can be shown that

$$\delta I=\frac{1}{1+\frac{R_2' E_1}{R(E_1-E_2)}}\left\{\frac{R_2' E_1}{R(E_1-E_2)}\,\delta E_1+\delta E_2-\left[1+\frac{R_2' E_1}{R(E_1-E_2)}\right]\delta R-\frac{R_2'}{R}\,\delta R_1\right\}; \tag{2}$$

$$I_2=\frac{1}{1+\frac{R_2' E_1}{R(E_1-E_2)}}\left[\frac{E_1}{E_1-E_2}(\delta E_1-\delta E_2)+\delta R-\delta R_1\right]\cdot I, \tag{3}$$

where δI is the relative variation of the stabilized current; δE_1, δE_2, δR_1 and δR are the relative variations of the emf and the resistances in the circuit, R_2' is the internal (real) resistance of the standard cell.

Let us assume that $\frac{R_2' E_1}{R(E_1-E_2)} \ll 1$, with the knowledge that δE_2, δR and δR_1 are usually one to two orders lower than δE_1; we can then write, approximately,

$$\delta I = \frac{R_2' E_1}{R(E_1 - E_2)} \delta E_1 + \delta E_2 - \delta R; \tag{4}$$

$$I_2 = \frac{E_1}{E_1 - E_2} \delta E_1 I. \tag{5}$$

The appearance of current I_2 produces in the standard cell a concentration polarization emf ΔE_2, which depends on the intensity and duration of the flowing current. Hence the relative variations of the standard cell emf δE_2 will be determined not only by the lapse of time and the effect of external factors, but also by the value of δE_1. In order to eliminate this concealed relation from (4) let us assume that

$$\Delta E_2 = R_2'' I_2, \tag{6}$$

where R_2'' is the polarization (apparent) resistance of the standard cell. The relative variation in the standard cell emf due to polarization will then be

$$\delta E_2' = \frac{\Delta E_2}{E_2} = \frac{R_2'' I_2}{RI} = \frac{R_2'' E_1}{R(E_1 - E_2)} \delta E_1. \tag{7}$$

Combining (4) and (7) we finally obtain

$$\delta I = \frac{(R_2' + R_2'') E_1}{R(E_1 - E_2)} \delta E_1 + \delta E_2 - \delta R. \tag{8}$$

We find the stabilization coefficient of the circuit from (8) by assuming that $\delta E_2 = \delta R = 0$:

$$K = \frac{\delta E_1}{\delta I} = \frac{R}{R_2' + R_2''}\left(1 - \frac{E_2}{E_1}\right) = \frac{R}{R_2}\left(1 - \frac{E_2}{E_1}\right), \tag{9}$$

where R_2 is the total resistance of the standard cell.

An improvement in the stabilizer parameters and a further simplification of the computing formulas are possible if $E_1 \gg E_2$. In practice it is not difficult to attain $E_1 = (5\text{-}10)E_2$. In this case

$$K \approx \frac{R}{R_2} = \frac{E_1}{IR_2}; \quad I_2 \approx I \delta E_1. \tag{10}$$

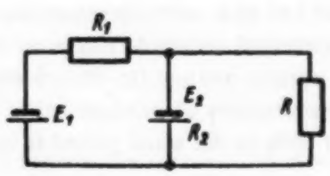

Fig. 1

The circuit in Fig. 1 is intended for the stabilization of one value of current I with a stabilization voltage equal to the standard cell emf (approximately 1.02 v). In order to stabilize several values of the current a switch is necessary which simultaneously changes resistors R and R_1 according to (1). It is also possible to raise the stabilization voltage by making up E_2 with two or more series-connected standard cells.

If stabilization according to Fig. 1 is insufficient it is possible to use a two-section circuit (Fig. 2) whose stabilization coefficient is

$$K = \frac{R^2}{R_2 R_3}\left(1 - \frac{E_2}{E_1}\right)\left(\frac{E_2}{E_3} - 1\right). \tag{11}$$

It is then obvious that $E_1 > E_2 > E_3$. If E_3 consists of a single standard cell, the first stabilizing source E_2 must consist of two or three standard cells connected in series.

The above relations complete the computation of a stabilizer circuit. Let us now examine the standard cell parameters E_2, δE_2 and R_2. These parameters are the main factors in determining the properties of a stabilizer, i. e., the possible values of I and δI.

Stability of a standard cell emf. For operation in a stabilizer circuit it is possible to use both saturated and unsaturated standard cells. The values of standard cell emfs and their permissible variations during one year are given in GOST (All-Union State Standard) 1954-55 and instruction 176-55.

The initial value of a standard cell emf which it is intended to use in a stabilizer circuit should be chosen in the range of 1.01858 to 1.01864 v for saturated, and in the range of 1.0189 to 1.0192 v for unsaturated standard cells. These ranges are three times narrower than those provided by specifications, thus ensuring to a certain extent a greater stability in the emf of the selected standard cells. Normally the emf of standard cells becomes more stable as it approaches the mean value (1.0186 v for saturated and 1.019 v for unsaturated standard cells). This property, for instance, serves for the preliminary selection of standard cells into reference groups. Moreover, only new standard cells a few months after manufacture should be used in stabilizers. It should be noted that under normal laboratory conditions, when no current is consumed the life of saturated standard cells amounts to 10-30 years, and that of unsaturated cells to about 5 years.

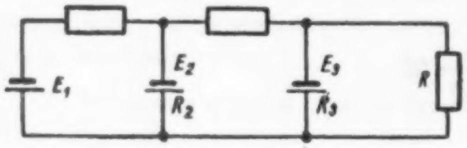

Fig. 2

Investigations of the stability of Soviet mass-produced standard cells carried out by D. I. Mendeleev VNIIM (All-Union Scientific Research Institute of Metrology) under the guidance of Prof. A. K. Kolosov have shown that the emf of saturated cells does not vary during one year by more than 20 μv*; that of unsaturated cells does not decrease by more than 100 μv. The standard cell emf tolerances specified for one year by the GOST are several times larger than the above figures. This circumstance is due to the fact that these tolerances take into account on the one hand the maximum permissible variations which meet the measurement techniques requirements, and on the other the possible effect of unfavorable conditions in the use of the cells (heavy vibrations, the use of current and accidental short circuits, sudden variations in temperature, operation at high temperatures in the case of unsaturated cells, etc.).

When the standard cells are operating under laboratory conditions in a stabilizer circuit it is possible tentatively to adopt the above values for their emfs.

The effect of temperature on the emf of unsaturated standard cells can be normally neglected, since their temperature coefficient does not exceed 5 μv/°C. In order to eliminate the effect of temperature on the emf of saturated cells it is necessary either to maintain the temperature of the cells constant by means of an automatic thermostat, or use an automatic temperature compensation device. These problems must be dealt with independently. We shall only note that with appropriate thermostatic control or compensation it is possible to neglect the temperature effect of these cells. Irrespective of the type of standard cells used they must be placed in a massive thermal screen which will provide within the standard cell jacket a uniform temperature with deviations not exceeding 0.01°C.

Internal resistance of standard cells. The value of standard cells' internal resistance R_i is mainly determined by the size of their jackets. In the recently manufactured Soviet standard cells of a normal design the internal resistance is of the order of 300 ohm. This value consists of the electrolyte resistance of 150 ohm with the remaining 150 ohm being due to the resistance of the cadmium sulfate crystal layer (in the saturated standard cell) or to the resistance of the crust separation which retains the ingredients of a standard cell in a certain position (in the unsaturated standard cells). In time the resistance of saturated standard cells rises slowly. In the majority of standard cells (about 60%) in 5 years' time the resistance rises by 100-200 ohm, whereas in some (about 10%) in the same period it is increased by a factor of 5-10, and sometimes even 20-30**.

The reduction (restoration) of the resistance to its original value is normally obtained by passing through the standard cell for a short time an audio-frequency current (20 v, 10,000 cps for 1 sec) [3]. This treatment does not eliminate the cause of the rise in the internal resistance; however, it reduces this resistance for a considerable time. It is also possible to lower the resistance by violent shaking of the standard cell. However, such treatment is difficult to regulate and it may have detrimental effects on the cell. The internal resistance of unsaturated cells usually remains constant with time.

*For the purpose of comparison let us note that the variation in the emf of reference standard cells made and kept at VNIIM does not exceed 2 μv per year.

**According to the test results of standard cells obtained by the VNIIM in 1956-1960.

The resistance of standard cells can be decreased to 50-100 ohm by increasing the surface of the electrodes (limb diameter of 20-25 mm). The resistance can also be reduced by connecting several standard cells in parallel. Such a connection does not have an adverse effect on the emf stability of the group providing the difference in the emfs of separate cells does not exceed 30 μ v for saturated and 100 μ v for unsaturated standard cells.

Polarization resistance of standard cells. The standard cells' polarization resistance R_2'' is found from the experimental relationship $\Delta E_2 = f(I, t)$. An analysis of the data obtained from literature [1-4] and the experiments carried out by the author at the VNIIM* lead to the following conclusions.

The polarization emf of newly produced Soviet standard cells amounts to 10 μ v per 1 coulomb for saturated and 20 μ v per 1 coulomb for unsaturated standard cells in the current range of 0.01 - 10 μ a. It will be seen from Fig. 3 that these data are strictly tentative; however, they are satisfactory for normal computations. The polarization emf of standard cells operated for a long time may, however, exceed the above values by a factor of 5-10 [6].

The polarization resistance of standard cells depends on the intensity and duration of the discharge current. This resistance decreases with a rising current (Fig. 4). Since standard cells are normally used in precision electrical measuring circuits intended for long-term operation (months, years), the discharge current should not exceed 1 μ a. Under such conditions the drop in the emf will amount to some 0.05%. Hence, polarization resistance rises in one year from a few tens to several hundred ohms. With a discharge current 0.1 μ a the drop in the emf for one year will amount to some 0.003%, and hence the polarization resistance will amount to 300 ohm. For currents smaller than 0.1 μ a the polarization resistance no longer depends on the current.

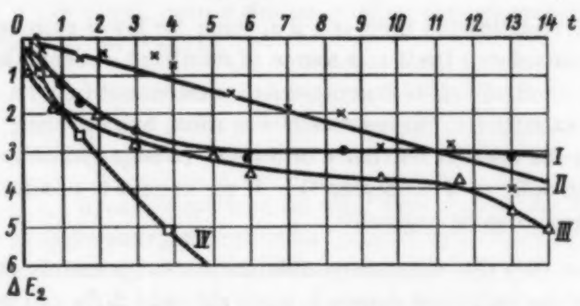

Fig. 3. Mean variations in standard cell emf (from 10 standard cells) under the effect of a continuous discharge current. I) Denoted by ●, current of 0.1 μ a, saturated standard cells, ΔE_2 in 10 μ v; II) denoted by ×, current of 1 μ a, saturated standard cells, ΔE_2 in 100 μ v; III) denoted by △, current of 1 μ a, unsaturated standard cells, ΔE_2 in 100 μ v; IV) denoted by □, current of 10 μ a, saturated standard cells, ΔE_2 in 1000 μ v.

Thus, the resistance of a standard cell with a continuous discharging for one year by a current of not more than 1 μ a can be tentatively estimated to amount to 400 ohm.

The value of the polarization resistance is determined by the effective surface of the electrodes. It has been established experimentally** that the polarization emf is due mainly to the positive electrode (about 0.9 ΔE_2). Hence the polarization resistance can be most effectively decreased by increasing the surface of this electrode.

The polarization of standard cells is considerably larger in charging than in discharging [3]. For a current of 1 μ a the charging polarization is ten times greater than the discharging polarization (Fig. 5). However, charging and discharging currents not exceeding 0.1 μ a have the same effect on the standard cell emf. It is, therefore, advisable to make the standard cells operate in the stabilizers in a discharging condition only***.

The above operating condition can be easily met if stable chemical sources of current are used for the stabilizer supplies, since the emf of such sources decreases with time. If, however, the source E_1 should consist of a stabilized rectifier fed from the mains, the standard cell should be artificially provided with a discharging operating condition, in order to meet requirement (1) for a maximum positive value of δE_1 instead of $\delta E_1 = 0$ as previously indicated. If the standard cell is fed from a rectifier it will inevitably pass an alternating current. It has been shown [3] than an alternating current not exceeding 10 μ a has no noticeable effect on the standard cell emf.

For the purpose of measuring the polarization resistance it is possible to use the circuit and technique described in the GOST and the instruction for measuring the internal resistance of standard cells. The polarization emf of a standard cell discharged for 5 minutes by a current of 0.1 μ a can serve as a criterion whether this cell is suitable

*A. S. Savushkina participated in this experiment.

**V. V. Myuller has produced special standard cells.

***The recommendation refers to standard cells operating as reference elements in compensation-type voltage stabilizers.

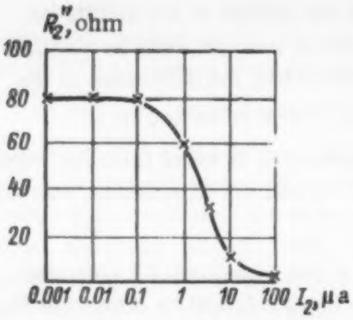

Fig. 4. Relation of the polarization resistance of an "average" standard cell to the discharge current. The discharging time amounts to 5 minutes.

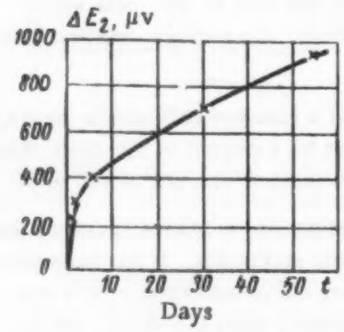

Fig. 5. Mean variation of standard cell emfs (of 3 standard cells) under the effect of a continuously charging current of 1 μ a.

for a stabilizer circuit. A polarization emf of 10 μ v (polarization resistance of 100 ohm) can be considered permissible. The measurement of the polarization emf, i. e., the difference of a standard cell emf before and after passing a current through it, should be made with an error not exceeding 1 μ v; the interval between the switching off of the current and the measurement of the standard cell emf should be as small as possible (5-10 sec).

Stabilizer parameters. The limiting values of the stabilized current are determined by the following considerations. Since the above stabilizer circuit is very simple and reliable it is desirable to make the source E_1 also simple and reliable. At the present time it is possible to recommend for E_1 either mercury oxide elements [7], which provide currents up to 100 μ a for a δE_1 of the order of several tenths of a percent, or silicon stabilizers [8], which provide a δ E_1 of the order of several hundredths of a percent. Assuming δ $E_1 = 0.03\%$ and $I_2 = 1$ μ a, we obtain according to (10) I = 3000 μ a. For $\delta E_2 = 0.3\%$ and $I_2 =$ 3 μ a (3 standard cells connected in parallel) we obtain I = 1000 μ a.

On the other hand, the smallest stabilization coefficient can be of the order 3-5. Assuming that K = 3 and $R_2 = 600$ ohm we shall find from (10) that R = 1500 ohm; hence current I = 600 μ a. For K = 3 and $R_2 = 200$ ohm (three standard cells connected in parallel) R = 600 ohm, hence current I = 1600 μ a. It can be considered, therefore, that the top current stabilization limit is equal to approximately 1000 μ a.

The lowest current stabilization limit is 1 μ a, since for lower currents it is better to use the standard cell itself as a source of stabilized current. In such applications of the standard cell it is necessary to consider not only its total resistance and the variation of this resistance with time, but also the effect of temperature on the internal resistance of the cell (a negative temperature coefficient of the order of 0.03 degree^{-1}). If the current is smaller than 0.1 μ a the latter factor can be neglected.

It will be seen from (10) that the stabilization coefficient is mainly determined by the value of the stabilized current I, since the ratio E_2/R_2 can be varied in narrow limits. For currents of the order of 100 μ a the stabilization coefficient is K ≈ 20, and for currents of the order of 10 μ a it is K ≈ 200.

However, the attainable stabilization accuracy is determined mainly by the instability of the standard cell. If the required stabilization accuracy δ I = 0.01-0.03 % it is advisable to use an unsaturated standard cell (without compensation); if however δ I = 0.003-0.01 %, a saturated standard cell should be used.

Conclusion. The sphere of application of the above circuit consists in long-term stabilization (months or years) of dc currents in the range of 1 to 1000 μ a with an error of 0.003-0.03%.

The application of the above stabilizers includes the stabilization of the working current of high-precision dc automatic compensators (with a digital display).

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APPLICABILITY OF KOTEL'NIKOV'S THEOREM TO DISCRETE MEASUREMENT TECHNIQUES

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One of the most important problems in the theory of analog-digital converters and digital instruments is the evaluation of the dynamic errors arising in measurements of continuously varying quantities.

The measurement of a variable quantity consists in determining its instantaneous values which correspond to given values of the independent variable. Normally time is used as the independent variable.

Discrepancies between the values of the variable and the time to which they are referred, and the error in determining the amplitude of the signal due to the variation of the measured variable during the measuring process, determine the dynamic error of the instrument.

Discrete measuring systems are subject to three types of dynamic errors:

dynamic errors of the first kind, due to a functional relation between the values of the input and output signals, which differs from a similar relation with respect to electrical measurements under static conditions. This problem is dealt with in [1];

dynamic errors of the second kind, due to the variation of the measured quantities during the discrete measuring process. One of the effective methods of dealing with such errors is the clamping of the input signal level during the discrete readout [2];

dynamic errors of the third kind, due to the discrete nature of measurements in digital instruments. The error in representing a continuous quantity by a series of its discrete values is a function of the time intervals Δt between the readings of the variable quantity.

Even if one assumes that the dynamic errors of the first and second kind do not exist and the instantaneous value of the function is determined accurately, there still remains the problem of ascertaining the values of the continuously varying quantity in the intervals between its measured values. Errors of this nature can be called errors of approximation. In this article attention is mainly paid to this problem.

In the theory of information the problem of a discrete transmission of a continuous signal is solved by means of a frequency representation of the signal. The solution is based on Kotel'nikov's theorem derived for finding the conditions for transmitting telephone communications by pulse methods [3].

The proof of Kotel'nikov's theorem is based on expanding function f(t) into a special kind of series according to which

$$f(t)=\sum_{-\infty}^{+\infty} f(k\Delta t)\,\frac{\sin\omega_s(t-k\Delta t)}{\omega_s(t-k\Delta t)}. \tag{1}$$

It follows from (1) that function f(t) is determined by its instantaneous values $f(k\Delta t)$, taken in intervals of

$$\Delta t=\frac{\pi}{\omega_s}=\frac{1}{2f_s},$$

where f_s is the highest frequency registered in the spectrum of the measured function.

It follows from this theorem that function f(t), which operates in interval T, can be fully expressed by its N discrete values:

$$N=\frac{T}{\Delta t}=2f_sT. \tag{2}$$

The applicability of this information theory proposition to the digital instrument circuits can be decided by analyzing the possibility of meeting a number of specific requirements of the measurement technique.

Firstly, it becomes impossible to use a low-pass filter at the output of the measuring circuit in order to provide an automatic summation of the signal components determined from (1) in the same manner as it is done in the communication technique.

TABLE 1

Error made in representing the measured signal, γ, %	1	0.5	0.1
Number of evaluated harmonic components, n	8	11	24
Number of measurements required per period according to Kotel'nikov's theorem, N.	16	22	48

TABLE 2

Error, $\pm$ k δ	0.5	1	3	5	10	50
Error, %	0.05	0.1	0.3	0.5	1	5
Transformation frequency, N/sec	705979.9	195405.1	22377	8077.8	2021	80.9

Secondly, the nature of the measured signal is not known before the measurement. With an equal degree of probability it can be either a complex or an almost harmonic signal. In the latter instance it is impossible to use (2) directly.

Thirdly, the approximation of the measured signal by means of a sum of sinusoidal functions of the type sin x/x is inconvenient for measurement applications, since additional computation is required for every point lying between the measured discrete values of the function.

Fourthly, the approximation of the relation under consideration by means of a sum of functions of the type of sin x or sin x/x for many signals of a complex form (for instance, rectangular or exponential) expands into a series whose convergence is very low. In order to represent the function with a given accuracy a much larger number of terms is required than for other types of approximations.

As has already been stated, it is impossible to use (2) when the signal under consideration is of an almost harmonic shape. In measuring such a signal at any frequency we shall always obtain according to Kotel'nikov's theorem only two discrete values for the function per period, by means of which it is impossible to reproduce the process unless its amplitude and frequency are known. In solving a measurement problem we must assume that both these values are unknown to us.

However, a detailed examination of this particular case shows that Kotel'nikov's theorem can be applied only if one assumes that measurements are made during a half-period of the signal's variations, i. e., in assuming that the signal has the form of a nonperiodic sine or cosine pulse.

The width of the function's spectrum is determined as the interval outside which the spectral density is smaller than a certain given value. It is obvious that with such an assumption a measurement error is made whose value is determined by the ratio of the evaluated and neglected harmonic components.

The above can be easily illustrated by means of the harmonic synthesis method [4]. According to this method the expansion into a Fourier series can be represented in a closed form, thus making it easy to determine the error made in representing the signal by means of a shortened Fourier series. Thus for a nonperiodic sinusoidal pulse [4, p. 153],

$$(x)=\sum_{n=1}^{\infty}\frac{1}{(2n-1)^3}\sin(2n-1)x=\frac{\pi^2}{8}x-\frac{\pi}{8}x^2 \qquad (3)$$

for $0<x<\pi$.

Formula (3) makes it possible to determine the error made in approximating the above function when it is represented by n terms of the series.

If we denote the fundamental frequency by f_1, then $f_s = nf_1$ and we shall obtain from (2) that N = 2n.

Further calculations by means of (3) for various values of n provide results given in Table 1.

A more general method of discovering the required expansion harmonics for a given accuracy of signal representation is given in [5].

The above method is based on the fact that the total signal energy P_s, which is determined by its spectral density $S(\omega)$, includes the error signal energy (P_e):

$$P_s = P + P_e. \tag{4}$$

The limiting frequency of the spectrum is obtained from (4) providing the total energy of the neglected harmonics in the spectrum does not exceed the error signal energy P_e.

The author of this article has derived the relation between the transformation frequency N and accuracy with which the function can be represented for an exponential signal with a time constant τ :

$$N = \frac{1}{\pi} \operatorname{tg} \frac{\pi}{2} \left(1 - \frac{k^2}{n^2}\right) \cdot \frac{1}{\tau}. \tag{5}$$

TABLE 3

Approximation error, %	Sinusoidal signal		Exponential signal	
	Number of measurements per period		Number of measurements per second	
	by Kotel'nikov's theorem	linear approximation	by Kotel'nikov's theorem	linear approximation
0.05	–	–	705980	20
0.1	48	67	195405	14
0.3	–	–	22377	8
0.5	22	31	8078	7
1.0	16	22	2021	4
5.0	–	–	81	3

The computing data for the case $\tau = 1$ sec, a maximum number of various levels $F = 999\delta$, and an error (in fractions) of the minimum discernible level $\pm k\delta$, are shown in Table 2.

Thus, in sinusoidal pulses the energy is concentrated in the first 18-20 harmonics, but for an exponential signal the energy falls with frequency very slowly and is not equal to zero over the entire frequency range of 0 to ∞. This explains such a striking difference in the number of required transformations for the same accuracy. Hence, approximation of the signal by means of a function of the form of sin $\underline{x}$ is suitable for the first example (16 transformations per period for an accuracy of 1%) and is unsuitable for an exponential signal (2000 transformations for $\tau = 1$ sec and an accuracy of 1%).

The above results become easily understandable if the shapes of the investigated signals and those of the approximating functions are compared. It is obvious that it is not expedient to approximate an exponential function by means of a sinusoidal curve. It is, therefore, necessary to find other shapes of approximating curves which would provide a considerable reduction in the required speed of approximation for exponential type functions.

The simplest form of approximating any function consists of representing it by means of straight-line segments. Discrete measurements of instantaneous values of function x(t) are made at given intervals of time Δt, which represents an approximation of the function x(t) by a broken line (see figure).

If a function x(t) which is continuous over the whole time interval during which its discrete values are registered, and which has in this interval a continuous first and second derivative, is approximated by means of a polynomial x'(t) and by the method of linear interpolation, we shall obtain the relationship [6]

$$N = \frac{1}{\Delta t_i} = \sqrt{\frac{\alpha_m}{8\delta}}, \tag{6}$$

where N is the frequency of discrete measurements, Δt_i are the time intervals at which the instantaneous values of the investigated signal are measured, α_m is the maximum value of the second derivative of the signal in the portion $t_n - t_0$, and δ is the given value of the approximation error.

Expression (6) makes it possible to determine frequency N of discrete measurements with respect to a given approximation error and the nature of the process under investigation.

When the linear interpolation method is applied to either of the previously examined signal shapes, the computation will provide the required rate of transformation for a given accuracy of approximation.

There exist other methods of approximating functions [7]. They include circular and parabolic interpolations.

In circular interpolation straight lines and arcs of a circle are used, the arcs being drawn in such a manner that they are tangential to the straight-line segments at the test points. The error in this method of approximation depends on the amplitude of the tested signal. Moreover, this method requires the computation of the approximating circle's radii.

In parabolic interpolation the arcs of circumferences are replaced by parabolas which are tangential at the points of discrete measurements to the straight lines connecting these points.

Under certain conditions this method provides the required error with a considerably lower rate of measurements than in the case of a linear approximation.

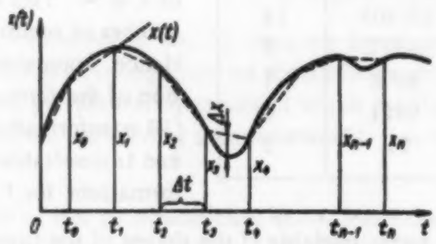
The error is independent of the signal amplitude, but the method is difficult to apply to signals of an arbitrary shape when it is necessary to determine the constants of the parabola at all the inflection points of the function. In such a case, it would appear, it is only practicable to apply a linear interpolation which is analyzed below.

For a sinusoidal signal $x = A \sin \omega t$ formula (6) can be transformed by inserting the values of

$$\alpha_m = |\omega^2 A|; \quad \gamma = \frac{\Delta A}{A} \cdot 100\,\%.$$

where A is the signal amplitude. Then

$$N = \sqrt{\frac{\omega^2 A}{8 \Delta A}} = \sqrt{\frac{(2\pi)^2 f^2 \cdot 100}{8\gamma}} = 22.2 \frac{f}{\sqrt{\gamma}} \text{ meas/sec.} \tag{7}$$



The required number of measurements per period will therefore be

$$N' = 22.2\, T \frac{f}{\sqrt{\gamma}} = \frac{22.2}{\sqrt{\gamma}}, \tag{8}$$

where γ is the approximation error, %.

In the case of an exponential signal

$$x = a e^{-\frac{t}{\tau}}; \quad \alpha_m = \frac{A}{\tau^2}; \quad \gamma = \frac{\Delta A}{A} \cdot 100\,\%, \tag{9}$$

where A is the maximum value of the signal, ΔA is the absolute value of the error.

$$N = \sqrt{\frac{100}{8\tau^2 \gamma}} \approx \frac{4}{\tau \sqrt{\gamma}} \text{ meas/sec.} \tag{10}$$

For the case of $\tau = 1$ sec

$$N = \frac{4}{\sqrt{\gamma}} \text{ meas/sec.} \tag{11}$$

A comparison of the results obtained from (8) and (11) is given in Table 3.

Conclusions. 1. The linear interpolation method requires a higher (by a factor of 1.4) rate of transformation for signals of the simplest form compared with approximation functions of the type of sin $\underline{x}$ or sin x/x, but it reduces that rate to hundredths and thousandths of that required for the latter functions in transforming variables of a more complex form, for instance, of the exponential type.

2. Kotel'nikov's theorem applies to approximations by means of a function of the type of sin x/x, hence it is not suitable for discrete measurements for the reasons given above; moreover, its application makes it difficult to obtain the values of the measured variable between two adjacent discrete readouts.

3. A linear approximation appears to be sufficiently suitable and convenient for discrete measurements of continuous variables; its application does not require additional computations, devices or gauges, which would have been required for any other approximating function.

4. When signals approaching a harmonic form are measured by the discrete method, the main propositions of Kotel'nikov's theorem hold if the highest frequency f_s of the measured signal spectrum is taken to be the harmonic frequency which must be evaluated in order to obtain a discrete representation of a harmonic pulse signal with a given accuracy.

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TRANSISTORIZED STABILIZERS FOR FEEDING TESTING INSTALLATIONS

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Very exacting requirements are specified for supplies to potentiometric installations used for checking precision electrical measuring instruments such as grade 0.5 and 0.2 ammeters, voltmeters and wattmeters. During measurements the variation in the supply current (voltage) must not exceed one-hundredth of a percent, and the supply voltage must not contain an ac component. Until recently this problem was solved by using relatively powerful storage batteries, which are available in any potentiometric testing installation.

In recent years electronic (tube) stabilizers have been developed for feeding parallel instrument circuits, thus replacing the cumbersome storage batteries with voltages up to 300 v. However, it is impossible to use electronic stabilizers for feeding series-connected (current) instruments which require sources of a few volts but currents rising up to several tens of amperes. Such stabilizers can be made but they would be too cumbersome and their efficiency extremely low.

Relatively high stability can be obtained from mains supplies by using semiconductor rectifiers connected to the mains through stepdown transformers and ac stabilizers. However, in such stabilized rectifiers it is in practice impossible to suppress the ac component in the output voltage, since the filters required for this purpose, owing to low frequencies and large currents, would have excessively large dimensions.

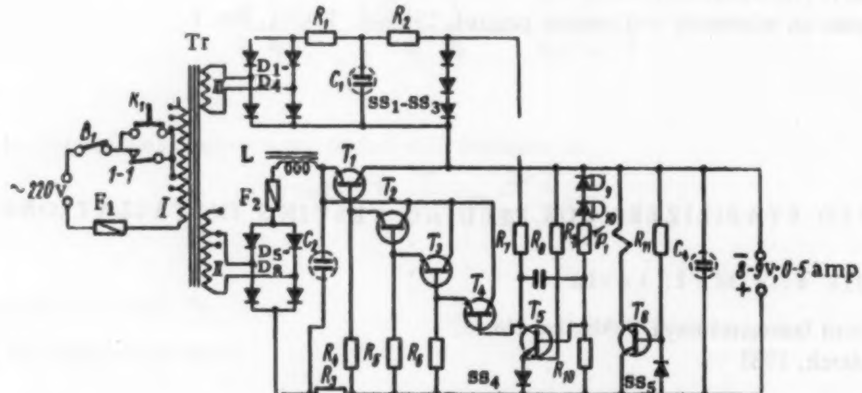
This is why to date storage batteries are used in all precision testing installations for feeding series-connected instruments.

The problem of providing such supplies can be solved fairly easily if semiconductor voltage stabilizers are used, with silicon stabilitrons as reference voltage sources, and the output voltage stabilized against variations in the supply voltage or load current by means of semiconductor triodes (transistors).

Modern transistor voltage and current stabilizers can provide for normal ac mains supplies a very high stability of the output voltage and current, a negligible ac component and high efficiency. Such stabilizers can be designed for voltages of 7-10 v and lower for currents in the range from a few milliamperes to tens and even hundreds of amperes.

The attached figure shows the schematic of a stabilizer with an output voltage of 8-9 v and a current up to 5 amp*. The principle of operation of this circuit is described in [1], the data for the components of this circuit are given in the attached specification.

The stabilizer employs a series transistor control which consists of a transistor combination T_1-T_4 with a single-stage transistor T_5 dc amplifier (DCA) and a silicon stabilitron SS_4, which is connected in the amplifier emitter circuit in order to provide a reference voltage. In order to raise the stability of the circuit the amplifier collector is fed from a separate parametric stabilizer SS_1-SS_3, which is connected in series with the output voltage. In order to protect the stabilizer from current overloads and short circuits a protection circuit is used consisting of transistor T_6, stabilitron SS_5, resistors R_3 and R_{11} and relay P_1. In a normal condition T_6 is conducting and relay P_1 contact 1-1 is closed. If the load current rises above a set value, which occurs in overloading or with a short circuit, the voltage drop across resistor R_3 (0.06 ohm) and hence across stabilitron SS_5, which is connected in a forward direction, increases. This leads to a rise in the stabilitron current and a decrease in the base current of transistor T_6 until the latter is completely blocked. This releases relay P_1 and its normally open contact 1-1 disconnects the stabilizer's supply circuit. In order to reconnect the stabilizer it is necessary to press push-button K_1 which shorts the relay contact.



The stabilizer uses a temperature compensation circuit [1].

The stabilizer is designed to work from 220 v 50 cps mains. With a mains voltage variation of ±10 % the output voltage does not change by more than 0.005%. For a load current variation of from 0 to 5 amp the output voltage changes by less than 0.025%. The amplitude of the output voltage ac component does not exceed 0.5 mv, which amounts to less than 0.01%. The use of a temperature compensating circuit makes the output voltage almost independent of temperature.

* The stabilizer was developed by S. D. Dodik.

SCHEMATIC SPECIFICATIONS

B_1 is a tumbler switch type TV1-2 or any other type;
K_1 is a push-button switch of any type;
P_1 is a relay type RSM-1 or any other type for a voltage below 10 v and a current smaller than 40 ma;
D_1-D_4 are diodes type D7G–D7Zh, DGTs-24–DGTs-27;
D_5-D_8 are diodes types D303-D305 or 15 diodes types D7A-D7Zh, DGTs-21–DGTs-24 connected in parallel;
D_9-D_{10} are diodes types D7A-D7Zh or preferably D808-D813 (the number of diodes and the current through them are determined in adjusting the thermal compensation as indicated in [1]);
SS_1 - SS_3 is a stabilitron type D810-D811;
SS_4 is a stabilitron type D808;
SS_5 is a stabilitron of any type (D808-D813);
C_1 is a capacitor type ÉGTs of 50 μf, 200 v or type KÉ of 50 μf, 200 v;
C_2 and C_4 are capacitors type ÉGTs of 2000 μf, 20 v, or type KÉ of 2000 μf, 20 v;
C_3 is a capacitor type KBGI of 0.1 μf, 200 v or any other 0.1 μf capacitor;
R_1 is a resistor type MLT-2 of 1.8 kilohm ±10%;
R_2 is a resistor type MLT-2 of 3.6 kilohm ± 10%;
R_3 is a 0.06 ohm resistor wound with manganin 1.2 mm wire;
R_4 is a resistor type MLT-2 of 150 ohm ±10%;
R_5 and R_6 are resistors type MLT-0.5 of 3.6 kilohm ±10%;
R_7 is a resistor type BLP-1 of 10 kilohm ±5% or a wire-wound resistor of any type;
R_8 is a resistor type BLP-1 of 130 ohm ±5 % or a wire-wound resistor of any type;
R_9 is an adjustable manganin wire-wound resistor of 100 ohm; it is used when an accurate adjustment of the output voltage is required;
R_{10} is a manganin wire-wound resistor of 200 ohm ±5%;
R_{11} is a resistor type MLT-0.5 of 2 kilohm ±10%, it is selected according to the type of the relay;
T_1 is a transistor type P207-P208 mounted on a blackened duralumin heat dissipator with an area of 1500-2000 cm², or 5 transistors type P4B-P4D in parallel with 0.5 ohm compensating resistors in each emitter mounted on blackened duralumin heat dissipators with an area of 300-400 cm²;
T_2 is a transistor type P4B-P4D mounted on a heat dissipator with an area of 200-250 cm²;
T_3, T_4, T_5, and T_6 are transistors types P13-P16, P25-P26;
F_1 is a 0.5 amp fuse;
F_2 is a 5 amp fuse;
Tr is a transformer type Sh-25×35; its winding I has 1040 turns of 0.6 mm PÉV wire, it has taps at turns 720, 800, 880 and 960; its winding II has 90 turns of 1.7 mm PÉV wire, with taps at turns 70 and 80; its winding III has 300 turns of 0.15 mm PÉV wire:
L is a choke coil type Sh-25 × 35 with an air gap of 1 mm and its volume filled by a winding of 1.7 mm PÉV wire; its core made of Sh-25 iron can be replaced by Sh-20 iron with an appropriate readjustment of the rectifier.

Experience gained in operating the above stabilizer for several months in the VNIIK (All-Union Scientific Research Institute of the Committee of Standards, Measures and Measuring Instruments) Laboratory showed that it provides suitable conditions for testing pointer current instruments by means of a potentiometric installation, and can replace storage batteries.

A similar circuit can be used for stabilizers with current ratings up to 10, 20 and 50 amp by simply increasing the number of parallel-connected controlling power transistors and rectifiers. Transistors types P207-P208 or P4 can be used as power regulators. For mains voltage variations of ±10% and an amplitude of input voltage pulses of the order of 0.5 v, one transistor type P4 or P207-P208 can supply without additional ventilation a current of 1.2 or 6 amp respectively. For this purpose transistors are mounted on blackened duralumin heat dissipators of an area of 100-150 cm² per 1 w of dissipated power. If the mains are already stabilized with an error of ±2%, the current obtained from the transistors for the same dimensions of the heat dissipator can be raised by 50%. If transistors P4 or P207-P208 are operated in parallel they should be balanced by means of resistances of the order of 0.5 or 0.1 ohm respectively in their emitter circuits.

For a stable operation the voltage between the collector and emitter of transistor T_1 for a maximum load current of 5 amp and a mains voltage of 220 v−10% (198 v) must be about 3-4 v; for a voltage of 220 v it should be about 5 v, and for that of 220 v+10% (242 v) about 7 v.

A higher stability of the output voltage can be obtained by using two- or three-stage dc amplifiers, and also by placing the stabilizer measuring elements (T_5, SS_4, D_9, D_{10}, R_9, and R_{10}) in a miniature semiconductor thermostat [2].

The mass production by our industry of transistorized stabilizers for checking current circuits will provide a considerable saving, free test laboratories from cumbersome storage batteries, and in certain instances raise the accuracy of measurements.

LITERATURE CITED

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2. E. Baldinger, W. Czaja, Ein einfacher Thermostat mit transistoren, Z. für angewandte Mathematik und Physik, 1958, v. IXa.

SIMPLIFIED COMPUTATION OF A SINUSOIDAL FERRORESONANT STABILIZER

V. P. Pozdnyakov

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The ferroresonant stabilizers produced by our industry are unsuitable for operation in circuits which require a sinusoidal voltage. The output curve of such stabilizers is heavily distorted by higher harmonics, the third harmonic amounting to 20-35%, the fifth to 7-12% and the seventh to 3-5%. According to the existing standard the harmonic content of a sinusoidal voltage must not exceed 5%. This condition is met if the saturated coil current is connected to two additional resonance filters tuned to the third and fifth harmonics. This is attained by the circuit shown in Fig. 1, which provides a high degree of stabilization with mains voltage variations of ±25% and does not require laminations of a special shape.

Accurate computations for ferroresonant stabilizers do not exist and those for sinusoidal stabilizers have not been touched upon in literature. We therefore provide for a sinusoidal stabilizer shown in Fig. 1 a simplified computation based on empirical formulas, only applicable to this stabilizer.

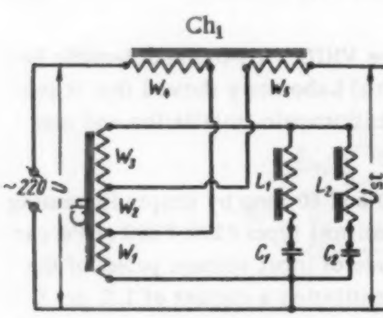

Fig. 1.

The stabilizer is required to work into a resistive load $P_L = 300$ w, at a mains voltage of U = 220 v, and a stabilized voltage of $U_{st} = 220$ v. Transformer iron type Sh-40, a cutout area $Q_0 = 12\ \text{cm}^2$ and a mean magnetic circuit length $l_m = 22.2$ cm should be used.

Saturated coil (of the autotransformer)

The core cross section $Q_c = 1.2\sqrt{P_L} = 21\ \text{cm}^2$. The number of turns in the windings is $W_1 + W_2 = 28\,U_{st}/Q_c = 293$, $W_1 = 18\,U/Q_c = 188$ and $W_2 = 105$ turns.

The total capacity of capacitors $C_1 + C_2$ is chosen according to the stabilizer power and the working voltage of the capacitors. Normally oiled-paper capacitors type SM-0.65-5 of 5 μf and a working voltage of 650 v are used in stabilizers. In this case ~5μf is taken for each 100w of power. Thus we obtain $C_1 + C_2 = 15$ μf.

The total number of turns in the coil will be

$$W_C = 1900\sqrt{\frac{l_m}{Q_C\,(C_1+C_2)}} = 503 \text{ turns.}$$

The number of turns in winding $W_3 = W_C - (W_1 + W_2) = 210$.

The voltage across the choke coil is

$$U_C = U_{st}\frac{W_C}{W_1 + W_2} = 377 \text{ v.}$$

Let us now check the total capacitance:

$$C_1+C_2=7500\ \frac{P_L}{U_C^2}=15\ \mu f.$$

Let us find the working voltage for which the capacitors must be calculated, $U_w \geq 1.6\ U_C = 603$ v.

We can thus choose the recommended capacitors.

<u>Unsaturated choke coil</u>

The Ch_1 core cross section is $Q_{C1} = 0.9 \cdot Q_C = 19\ cm^2$.

The number of turns in the main winding is $W_4 = 36\ U/Q_{C1} = 416$ turns.

The number of turns in the compensation winding is $W_5 = 0.22 \cdot W_4 = 91$ turns.

The unsaturated choke coil is made with an air gap of 2 mm. The size of the air gap is finally fixed in adjusting the stabilizer.

The diameters of wires used in various windings, including their enamel insulation, are: for winding W_1 $1.1\sqrt{I_z} = 1.28$ mm; W_2 $\sqrt{I_z} = 1.16$ mm; W_3 $0.9\sqrt{I_z} = 1.04$ mm; W_4 $1.05\sqrt{I_z} = 1.22$ mm; and W_5 $0.7\sqrt{I_z} = 0.95$ mm.

For purposes of adjustment, all the windings with the exception of W_4 have 4-6 tappings in regular intervals in the last 10% of their turns.

<u>Adjustment.</u> The initial adjustment of the stabilizer is made without coils L_1 and L_2 for a nominal load. For this adjustment the stabilized voltage should be 5-6% below the nominal.

A normally tuned stabilizer maintains the output voltage with an error of ±(0.5-0.7) % for mains voltage variations of ±10% and ±(1-1.5)% for variations of ±25% with a heavily distorted sine wave.

In the following operation the capacitors are divided in such a manner that the capacitance used for tuning the third harmonic filter is at least twice as large as that used for tuning the fifth harmonic filter.

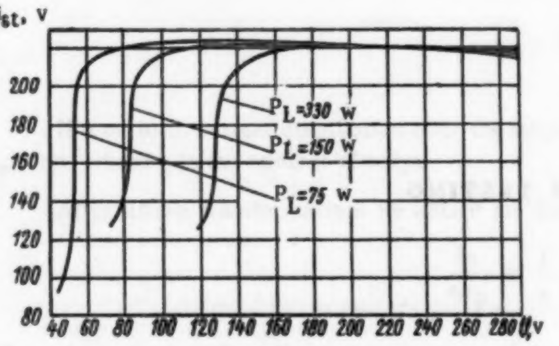

Fig. 2.

In our example the capacitance of the third harmonic filter is 10 μf and that of the fifth harmonic filter 5 μf.

The natural frequency for each resonant filter is determined from the basic relationship $f = 1/2\pi\sqrt{LC}$.

Since the values of capacitances are known we only need to find inductances for each filter. For the third harmonic filter it will be $L_1 = 1/\ 6.28^2 f^2 C = 0.11$ h, and for the fifth harmonic filter we find similarly $L_2 = 0.08$ h.

We can now find the approximate value of the effective currents in the capacitative branches for the fundamental frequency of 50 cps. For this purpose let us determine the current through one capacitor:

$$I_C = U_C\ 2\pi f C = 0.6\ \text{amp}.$$

Let us now find the filter currents for the third harmonic: $I_{f_3} = 1.25 \cdot 2\ I_C = 1.5$ amp, and for the fifth harmonic: $I_{f_5} = 1.15 \cdot I_C = 0.69$ amp.

The core cross sections for coils L_1 and L_2 will be for the third harmonic $Q_{st1} = (350 \cdot I_{f_3}^2 L_1)/Q_0 = 7.2\ cm^2$, and for the fifth harmonic we find similarly $Q_{st2} = 1.1\ cm^2$.

The air gaps in the coils for the third harmonic will be $l_{c1} = 5 I_{f_3}^2 L_1/Q_{st1} = 1.7$ mm; and for the fifth harmonic similarly $l_{c2} = 1.7$ mm.

The number of turns in the coils for the third harmonic will be $W_{L_1} = 8200 \sqrt{L_1 l_{c_1} / Q_{st_1}} = 418$ turns, and for the fifth harmonic similarly $W_{L_2} = 902$ turns.

The diameters of the winding wires will be $d_1 = 0.7 \sqrt{I_{f_3}} = 0.86$ mm, and similarly we find $d_2 = 0.58$ mm.

It is advisable to make the final adjustment of the stabilizer with an actual load and a cathode-ray oscilloscope.

By varying the air gaps in the coils the best sine wave is obtained, which should appear to be almost exactly the same as the mains voltage sine wave.

When the stabilizer operates with filters its output voltage rises to the nominal value and for ±10% variations of the mains voltage the stabilized voltage changes by ±(0.15-0.2)%, which is in fact difficult to obtain from a non-sinusoidal stabilizer.

The stabilized voltage is almost independent of the loading and mains voltage variations; however, dips in the output voltage at small loads occur at mains voltages only 25% below the nominal (Fig. 2).

With an ambient temperature of 20° such a stabilizer will withstand a prolonged overloading by 20%. If a resistive load is replaced by an inductive load the stabilizer characteristics are changed very little.

The efficiency of the stabilizer is relatively high and amounts to 85%.

In order to be completely satisfied with the operation of the stabilizer the percentage content of higher harmonics should be checked. This can be determined with sufficient accuracy for all practical purposes in the following manner.

The voltage terminals of an electrodynamic wattmeter are connected to an audio-frequency oscillator type ZG-10, and its current terminals to the stabilizer load. The oscillator is then set at 50 cps and the wattmeter readings noted. Next, without changing the current or voltage in the wattmeter the oscillator is set to the required harmonic and the maximum reading is again noted.

If the first reading is taken as 100% the remaining readings will correspond to the percentage harmonic content.

Conclusions. The above sinusoidal stabilizer is suitable for almost any devices used in checking and repairing electrical measuring instruments. The use of such a stabilizer frees the operator from the inconvenience of maintaining a given operating condition.

COMPUTATION AND DESIGN OF AC SHUNTS FOR TESTING COMMUTATION EQUIPMENT

I. B. Bolotin

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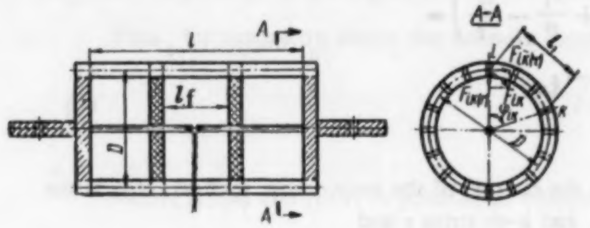
When testing electrical equipment for its commutation properties, dynamic stability, etc., it is necessary to measure among other quantities the short-circuit current, whose value reaches several hundred thousand amperes. The measurement of such large alternating currents with the required accuracy, especially in transient conditions, is very difficult, since to date there are no simple and reliable measuring instruments which would measure with the same accuracy both periodic and aperiodic components of a short-circuit current.

It is known that a normal current transformer with an iron core, which provides satisfactory measurements under steady state conditions, produces in a transient state large distortions due to the magnetization of its core by the aperiodic component of the current and to the existence of residual magnetism.

If a current transformer without an iron core is used (an air-cored transformer) its secondary winding will provide a voltage proportional to the derivative of the primary current. Hence, in order to reproduce the original current

it is necessary to integrate this voltage by means of an integrating device, thus leading to a more complicated measuring method and to a certain increase in the error of measurement.



The use of air-cored transformers with integrating devices is expedient for measuring large currents in high-tension circuits where the measuring circuit has to be isolated from the power circuit. When measuring large currents in a circuit which has fixed points with low potentials it is preferable to use ac shunts.

Considerable attention has been paid to the problem of measuring alternating currents by means of shunts at the Leningrad branch of the V. I. Lenin All-Union Electrotechnical Institute, by a team under the leadership of N. N. Nikiforovskii. It was established that measurements under short-circuiting conditions can be best made by cylindrical shunts with axial potential leadout conductors made in the shape of manganin strips placed in radial planes (see figure). This design has advantages compared with others, since it is very rigid and withstands electrodynamic effects, and provides an even distribution of current along the circumference (through the plates), since the resistance of a manganin strip is much larger than the transition resistance at the point where the strip is fixed to the butt. Thus the equivalent inductance of the shunt may be reduced to a very small value. In practice it is sufficient to make the ratio of the shunt length to the diameter of its butts approximately equal to 2.

For practical computation of an arc shunt intended for testing commutation equipment it is necessary to take into consideration that in operation the shunt is subjected to electrodynamic and thermal effects of the current flowing through it.

The formulas relating the electrodynamic forces and the electrical heating up of the current-conducting element with its geometrical dimensions and the current flowing through it are well known. On the basis of these formulas it is possible to derive convenient relations for the calculation of the shunt and to simplify them considerably.

<u>Computation of the shunt dc resistance.</u> The required total resistance r_s of the shunt is determined by the nominal current I_n for which the shunt is calculated and the voltage U_n across its voltage terminal which it is required to have for this current:

$$r_s = \frac{U_n}{I_n}.$$

If a uniform current distribution over the strips is assumed, the resistance r_p of each strip can be represented as $r_p = r_s n$, where n is the number of strips.

After simple transformations we obtain the design formula

$$\frac{l_p}{n q_p} = \frac{1}{\rho}\cdot\frac{U_n}{I_n} = 2.18\cdot 10^4\,\frac{U_n}{I_n}, \tag{1}$$

where l_p is the length of the manganin strips of the shunt, cm; q_p is the cross section of a strip, cm^2; ρ is the resistivity of manganin, assumed to be $0.46\cdot 10^{-4}$ ohm · cm; U_n is the nominal voltage drop, v; I_n is the current, amp.

<u>Computation of the shunt on the basis of its electrodynamic strength.</u> In this computation it is necessary to determine the force F_i acting in each strip of the shunt. In order to simplify the calculation the butt ends' effect is not considered, which only leads to a certain reserve strength.

When a current flows through the ith and kth strips, the whole length l of the ith strip is acted upon by force F_{lik}, which is equal to (see figure)

$$F_{lik} = 2.04\,\frac{I_i I_k}{l_1}\cdot 10^{-8}\left(\sqrt{l^2+l_1^2}-l_1\right),$$

where I_i and I_k are the effective currents in each strip, amp; l_1 is the distance between the strips, cm; l is the length of the strip, cm; F_{lik} is the force, kg-wt.

The mean value of the force acting per unit length F_{ik} can be represented in the form

$$F_{ik}=2.04\frac{I_iI_k}{l_i}\cdot 10^{-8}\left(\sqrt{1+\frac{l_i^2}{l^2}}-\frac{l_i}{l}\right)=$$

$$=2.04\frac{I_iI_k}{D\sin\frac{\varphi_{ik}}{2}}\cdot 10^{-8}\,\xi_{ik}, \tag{2}$$

where D is the diameter of the circumference which runs through the center of the strip cross section; φ_{ik} is the angle between the radii corresponding to the positions of the i-th and k-th strips; and

$$\xi_{ik}=\sqrt{1+\frac{l_i^2}{l^2}}-\frac{l_i}{l}.$$

The force F_{ik} can be resolved into its radial $F_{ik(r)}$ and its tangential $F_{ik(\tau)}$ components. The tangential components of the i-th strip,due to its interaction with the two adjacent strips which are symmetrical with respect to it,cancel each other.

The radial component is determined from the expression

$$F_{ik\,(r)}=F_{ik}\sin\frac{\varphi_{ik}}{2}=\frac{2.04\,I_iI_k}{D}\cdot 10^{-8}\,\xi_{ik}.$$

Since $I_i = I_k = I/n$, we have

$$F_{ik\,(r)}=\frac{2.04I^2}{Dn^2}\cdot 10^{-8}\,\xi_{ik}. \tag{3}$$

In this formula only coefficient ξ_{ik} depends on the ordinal of the strip.

The total force acting per unit length of each strip when current flows in all the strips of the shunt is directed radially toward the center and is equal to

$$F_i=\frac{2.04}{Dn^2}\cdot 10^{-8}\sum_{k=1}^{k=n-1}\xi_{ik}.$$

It can be shown that ξ_{ik} varies for an actual shunt in the limits of 0.7-1. Hence we can write

$$\sum_{k=1}^{k=n-1}\xi_{ik}=(n-1)\,\xi,$$

where ξ is a little smaller than 1.

In order to simplify the design formula let us assume that ξ = 1, which will lead to a certain increase in force F_i and hence to an additional reserve strength in the computation.

The design formula will then assume the form

$$F_i=\frac{2.04I^2}{D}\cdot\frac{n-1}{n^2}\cdot 10^{-8}. \tag{4}$$

Formula (4) relates the nominal current with the diameter and the required number of strips.

Calculation of the shunt on the basis of its electrothermal strength. In order to calculate the shunt on the basis of its electrothermal strength, considering the heating up to be adiabatic, let us use the formula which expresses the relation between the geometrical dimensions of the strips and the permissible current and heating-up temperature:

$$I=S\sqrt{\frac{c\gamma(T-T_0)}{0.24\rho t}},$$

where S is the total cross section of all the strips in the shunt, cm^2 ; c is the specific heat of manganin equal to 0.095 cal/g · deg; γ is the specific gravity of manganin, equal to 8.3 g/cm^3 ; $T-T_0$ is the temperature change of the shunt in t sec, °C; ρ is the resistivity of manganin equal to $0.46 \cdot 10^{-4}$ ohm-cm.

Thus, for manganin shunts the formula assumes the form

$$I = 268 S \sqrt{\frac{T-T_0}{t}}. \qquad (5)$$

From (1), (4) and (5) it is possible to compute the shunt.

Example. It is necessary to compute a shunt for I_n = 1000 amp, U_n = 100 mv, and a limiting current of 30 ka flowing during 0.3 sec.

From (5) we can determine the total cross section of all the strips of the shunt, assuming the permissible temperature rise in the manganin to be $T-T_0$ = 150°C :

$$S = \frac{30\,000}{268}\sqrt{\frac{0.3}{150}} = 5\ \text{cm}^2.$$

From (1) we find the length of the manganin strip, between the butt ends:

$$l_p = 5 \cdot 2.18 \cdot 10^4 \cdot 0.1 \cdot 10^{-3} = 10.9\ \text{cm}.$$

The total length l_s of the shunt is obtained by adding the thickness of the two butt ends. From constructional considerations and for a good heat dissipation let us take the thickness of the butt end as 1.5 cm; then l_s = 13.9 cm.

In order to obtain the smallest possible inductance of the shunt let us take its diameter as 7 cm.

The number of plates is chosen on the basis of cross section q_p of the available strips. Moreover, in order to ensure an adequate current distribution over the circumference the number of strips should not be less than 20. If we use strips with dimensions b = 0.1 cm and h = 1.0 cm, their cross section will be q_p = 0.1 cm^2 and their number n = S/q_p = 50.

In order to improve the mechanical strength of the shunt textolite struts 5 mm thick are placed in parallel with the butt ends so that the free span of the strips between the struts and the butt ends is l_f = 3.3 cm.

The force acting per unit length when a current of 30 ka is flowing can be found from (4):

$$F_l = \frac{2.04 \cdot 30\,000^2}{7} \cdot \frac{50-1}{50^2} \cdot 10^{-8} = 0.0525\ \text{kg-wt/cm}.$$

It can be considered with great accuracy that the force F_l acting in the middle span is uniformly distributed along the length of the strip. According to the well-known resistance-of-material formulas let us find the mechanical stresses in the strip, considering it as a beam fixed at both ends. The maximum bending moment will be $M_{max} = F_l l_f^2/12$ = 0.0476 kg-wt · cm.

The moment of resistance of the strip cross section $W = bh^2/6 = 0.0167$ cm^3. Tension $\sigma = M_{max}/W$ = 2.86 kg-wt/cm^2, which is well below the permissible value.

Let us now determine the limiting current which the shunt can withstand from the point of view of its electrodynamic strength. The tensile strength of manganin is 5000 kg-wt/cm^2.

Assuming the safety factor to be 5, we obtain the permissible stress as σ_p = 1000 kg-wt/cm^2.

$M_{max.p} = \sigma_p W$ = 16.7 kg-wt · cm; F_{lp} = 18.3 kg-wt/cm.

The maximum permissible current is

$$I_p = \sqrt{\frac{F_{lp}}{2.04} \cdot \frac{Dn^2}{(n-1)} \cdot 10^8} = 565\ \text{ka}.$$

The maximum permissible current amplitude is $I_{max.p}$ = 800 ka.

This figure is somewhat lower than the actual value owing to the values assumed and neglected.

Production characteristics of the shunt and test results. On the basis of similar computations a shunt was made for I_n = 1000 amp, U_n = 120 mv and a limiting current of 30 ka flowing for 0.3 sec. The manganin strips were placed in appropriate slots made in the copper butts and silver-soldered, thus providing a low contact transition resistance between copper and manganin.

Since in silver soldering the butts were raised to a temperature considerably exceeding the safe temperature for manganin, the shunt was placed during soldering in water in such a manner that all of the manganin was under water with the butt alone remaining above its surface. Hence, even if the water was heated up to its boiling point, the temperature of manganin could not exceed 100°C.

Since the resistance of one manganin strip came to about 6000 μ ohm and the transition resistance of the silver-soldered contact amounted to some 4-5 μ ohm, any discrepancies in the contact transition resistances of various strips can be neglected, and it can be assumed that the current distribution among the strips is uniform.

The errors of the shunt were determined in the following manner.

The shunt is designed to measure short-circuit currents which can be represented by the following expression:

$$i = I_m \sin(\omega t + \psi) - I \sin \psi e^{-\frac{t}{T_1}}, \tag{6}$$

where i and I_m are the instantaneous and amplitude values of the current; ψ is the switching phase of the current; T_1 is the decay time constant of the aperiodic current component.

The instantaneous voltage across the voltage terminals of the shunt can be expressed by the formula

$$\begin{aligned} U_s &= I_m r_s \sqrt{1+(\omega T_s)^2} \sin(\omega t + \psi - \varphi) - \\ &- I_m r_s \left(1 - \frac{T_s}{T_1}\right) \sin \psi e^{-t/T_1}, \end{aligned} \tag{7}$$

where $T_s = L_s/r_s$ is the equivalent time constant of the shunt; $\varphi = \tan^{-1} \omega T_s$ is the lag angle of the periodic component; r_s is the resistance of the shunt; L_s is the equivalent inductance of the shunt.

It will be seen from (7) that the proportionality between U_s and i is disrupted by coefficient $\sqrt{1+(\omega T_s)^2}$ in front of the periodic component and by angle φ, as well as by coefficient $1 - (T_s/T_1)$ in front of the aperiodic component in (7).

An analysis of (7) shows that in order to keep the error of reproducing both short-circuit components below 0.5% even in the case of complete aperiodicity, it is sufficient to make the time constant of the shunt meet the following requirement:

$$T_s \leq 0.005\ T_1. \tag{8}$$

Since in testing the equipment, the value of T_1 is normally in the range of 0.01-0.2 sec, it is necessary in order to satisfy (8) to make $T_s \leq 5 \cdot 10^{-5}$ sec.

At the frequency of 50 cps this condition can be replaced by the following:

$$\omega T_s \leq 1.57 \cdot 10^{-2}. \tag{9}$$

Conclusions. The shunt errors were measured on an ac potentiometer type R-56. The voltage terminals of the shunt were connected to the measuring terminals by means of a twisted pair. These tests led to the following conclusions:

the resistance of the shunt amounts to 119.6 μ ohm, which differs from the calculated value only by 0.33%;
the time constant of the shunt amounts to $\omega T_s = 0.835 \cdot 10^{-2}$, which satisfies (9).

It is possible to conclude from the measurement results that the shunt can be used for measuring short-circuit currents with an error not exceeding 0.5%.

The shunt was tested for thermal stability by passing through it 30 ka for 0.3 sec. The rise of temperature in this test did not exceed 100°C, which is below the calculated value. This circumstance is due to the fact that the

shunt is short and a considerable effect is produced by the dissipation of heat away from the manganin strips in the butts and massive busbars.

The shunt was extensively used in the commutation equipment laboratory of the Leningrad branch of the V. I. Lenin All-Union Electrotechnical Institute in testing high-tension equipment with satisfactory results.

TRANSISTORIZED DC INSTRUMENT AMPLIFIERS

V. V. Pavlov

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The use of semiconductor elements in dc instrument amplifiers has great possibilities, since it raises considerably their reliability, reduces the supply power and the size of the amplifiers.

The basic problem in designing transistorized dc instrument amplifiers consists in eliminating zero drift, which is especially difficult in direct-coupled amplifiers. The zero point of such amplifiers is determined by the reverse current of the collector junction, which is very sensitive to ambient temperature variations. With a rising temperature this current increases exponentially and doubles approximately every 11°C for germanium transistors and every 5.5°C for silicon transistors.

In direct-coupled amplifiers drift is decreased by means of balancing circuits, compensating feedbacks and individual thermal components such as thermistors, diodes, and other similar elements. However, all these methods require individual adjustments and are therefore unsuitable for commercial measuring instruments. Such amplifiers can only be used periodically, when they can be adjusted before each measurement as, for instance, under laboratory conditions.

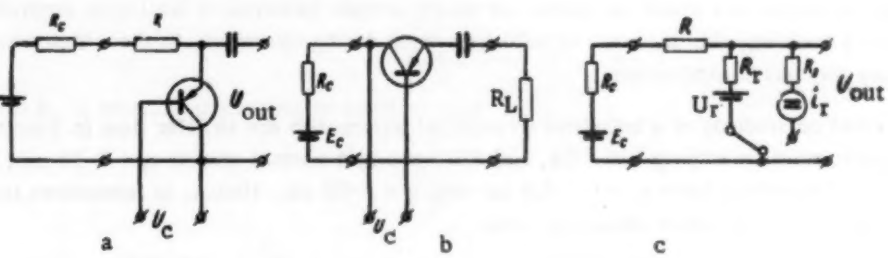

Fig. 1.

It is known that in commercial measuring equipment it is better to use amplifiers with a conversion of direct to alternating current and subsequent ac amplification. Such amplifiers must contain a dc-to-ac converter, a source of converting voltage, an ac amplifier and a rectifier. In modern measuring equipment vibrating inverters are used for dc to ac conversion. However, contactless inverters using semiconductor (diode or triode) or Hall elements are in many respects preferable to the former.

Investigation of the work of inverters using germanium and silicon diodes of various types (D2, D7, D201, D206 and others) has shown that their minimum sensitivity is equal to 3-5 mv. Their conversion factor, i. e., the ratio of the voltage drop at the output to the input dc voltage, amounts to 0.5-0.7. The conversion voltage may have either a sinusoidal or rectangular shape. Germanium diode inverters have unsatisfactory temperature characteristics. Their voltage drift in 8 hours of operation for temperature variations only in the room temperature range amounts on an average to 0.8 mv. Silicon diode inverters work satisfactorily with temperature variations of 45°C providing a voltage drift within 0.1 mv and a current drift within 10^{-10} amp (for 8 hours).

Lower dc voltages can be converted by means of Hall effect inverters. Their minimum sensitivity lies in the area of 50 μ v [1]. The conversion factor of Hall elements is determined from the formula

$$K_c = \frac{\mu B b}{l} \cdot 10^{-8}, \tag{1}$$

where μ is the carrier mobility, $cm^2/v \cdot sec$; B is the magnetic induction produced by the converting voltage, gauss; b is the width of the element plate, cm; l is the length of the plate, cm.

For instance, an antimonous indium Hall inverter would have $\mu = 2 \cdot 10^4$ $cm^2/v \cdot sec$, $l = 1$ cm, and b = 0.5 cm in a field with a magnetic induction of B = 2000 gauss and $K_c = 0.4$. A dc instrument amplifier circuit using a Hall element with a $K_c = 0.17$ is described in [1].

An essential defect of such an amplifier consists, in addition to its low conversion factor, in its temperature instability. Temperature compensation requires careful individual adjustment and in general is not satisfactory. Therefore Hall inverters are at present only used in laboratory dc amplifiers, where zero adjustments can be made before each measurement.

Inverters of weak dc signals operating with triodes in a switching condition are of considerable interest. In such inverters a converting sinusoidal or rectangular voltage is supplied between the base and the emitter or the collector of the transistor. And the emitter and collector electrodes are connected to the measured dc signal (see Figure 1a). Under the effect of the conversion voltage the base current varies from 0 to $i_{b\ max}$. If the input dc voltage is 0 no voltage will appear at the output. When a dc voltage is impressed on the input the triode will operate in a switching condition, it will be alternately blocked (collector current ≈ 0, collector voltage U_{co} maximum), and conducting (i_{co} maximum, $U_{co} \approx 0$). This will produce at the transistor output a voltage with an almost rectangular waveform proportional in amplitude to the dc voltage. If a dipolar conversion voltage is used the inverter will respond to a dc voltage of either polarity.

In this circuit the transistor can be connected either in parallel or in series with the source of the measured voltage (transducer). According to this the inverters are divided into those of a parallel and series type.

The smallest dc voltage which can be converted by means of such inverters is limited by the transistor's residual voltage and current parameters (u_r, i_r), whose value is determined by the intersection point of the transistor output characteristic zero branches. The levels of the residual parameters are higher than those of the internal noises of a transistor, making the amplifiers which use them less sensitive than balanced or Hall type amplifiers. However, their advantages consist in avoiding the necessity of adjusting them during operation, in their high conversion factor, their simplicity and easy thermal compensation.

With a reversed connection of a transistor its residual parameters are smaller than in a normal common emitter circuit. For instance, transistors types P13, P14, and P16 have in a normal circuit u_r = 3-10 mv, i_r = 10-20 μa; these triodes in a reversed connection have u_r = 0.3-0.8 mv and i_r = 5-10 μa. Hence, in instrument inverters of small dc signals reversed transistor connections should be used.

In Fig. 1, a and b circuits of parallel and series type inverters are shown with their transistors in a reversed connection. Fig. 1c shows the equivalent circuit of a transistor inverter of a parallel type with its residual parameters. In the conducting half-period of the transistor there remain a certain voltage, shown on the equivalent circuit in the form of a voltage source u_r, and a residual resistance R_r. During the blocking half-period, when the transistor internal resistance is large (R_3), there will be a residual current, shown on the equivalent circuit in the form of a current generator i_r.

The above equivalent circuit only holds for relatively low frequency inversion, since it does not take into consideration the transistor reactive parameters and, hence, the variation of the conversion with frequency. For inverters shown in Fig. 1 such an equivalent circuit holds for $f_c \leq 0.01 f_\alpha$ (f_α is the transistor's limiting frequency with respect to α). For instance, for transistors type P13, P14, and P16 f_c does not exceed 10-15 kc. For higher frequencies it is necessary to take into consideration variation of transistor parameters with frequency, which makes the computation much more complicated.

Let us now analyze an equivalent circuit for a transistorized inverter*. During the conducting half-period

*In the experiments for the inverter given below the load resistance, i.e., the input resistance of the amplifier (about 1300 ohms) is omitted, which, however, does not appreciably affect the accuracy of computations for $R_1 \leq 100$ ohm and $R_3 > 100$ kilohm.

the transistor passes a current due to the measured voltage and voltage u_r:

$$i_{cnd} = \frac{E_c + u_r}{R_c + R + R_r}; \tag{2}$$

the voltage across the output of the inverter during the conducting half-period is

$$U_{cnd} \approx E_c - i_{cnd}(R_c + R) = \frac{E_c}{(R_c+R)/R_r + 1} - \frac{u_r}{R_r/(R_c+R)+1}; \tag{3}$$

the voltage across the output of the inverter during the nonconducting half-period is

$$U_{non} \approx E_c + i_r \, (R_c + R), \tag{4}$$

providing $R_r \ll (R_C + R) \ll R_s$.

The voltage drop at the output of the inverter is

$$U_{out} = U_{non} - U_{cnd} = E_c + i_r \, (R_c+R) - \frac{E_c}{(R_c+R)/R_r + 1} + \frac{u_r}{R_r/(R_c+R)+1}; \tag{5}$$

the conversion factor of the circuit is

$$K_c = \frac{U_{out}}{E_c} = \frac{1}{R_r/(R_c+R)+1} + \frac{1}{E_c}\left[\frac{u_r}{R_r/(R_c+R)+1} + i_r \, (R_c+R)\right]. \tag{6}$$

For small values of i_r and u_r it is possible to consider that the second term in (6) is considerably smaller than the first and thus simplify considerably the formula for the conversion coefficient:

$$K_c = \frac{1}{\frac{R_r}{R_c+R}+1}. \tag{7}$$

In a particular case if R_C is small it is possible to consider that

$$K_C = \frac{1}{\frac{R_r}{R}+1}. \tag{8}$$

Expressions (5), (6) and (7) thus derived make it possible to judge the basic characteristics of transistorized inverters. It will be seen from (6) that the linearity of conversion is limited by the second term. In order to raise the linearity transistors with small parameters $(u, i, R)_r$ should be used. It follows from (6) and (7) that $K_C \to 1$, $U_{out} \to E_C$ when $R_r \to 0$; and $K_C \to 0$, $U_{out} \to 0$ when $(R_C + R) \to 0$. Thus for a maximum conversion factor $(K_C)_{max}$ it is necessary to use transistors with a high R_s and a small R_r, i. e., transistors of a switching type. It also follows from (7) that in order to raise K_C one should increase $(R_C + R)$; on the other hand it follows from (4) that in raising $(R_C + R)$ the residual voltage at the output of the circuit increases during the nonconducting half-period. For instance, if $i_r = 1 \mu a$, then for $(R_C + R) = 100$ ohm we obtain $(R_C + R)i_r = 0.1$ mv, for $(R_C + R) = 500$ ohm we have $(R_C + R)i_r = 0.5$ mv, and for $(R_C + R) = 1$ kilohm we have $(R_C + R)i_r = 1$ mv. Thus, for converting dc signals of 0-15 mv the value of $(R_C + R)$ must be limited to 100-200 ohm.

It follows from (4) that the residual voltages at the output of the inverter during the nonconducting half-period depend on the value of i_r. It should be noted that i_r rises exponentially with ambient temperature, thus increasing disproportionately the residual voltage at the inverter output. In order to decrease the residual voltage $(R_C + R)i_r$, it is necessary to connect R_s in parallel with the inverter output (Fig. 2a). The equivalent circuit of an inverter with a shunting resistor is shown in Fig. 2b.

The residual voltage at the output of the inverter will then decrease to

$$\frac{(R_c+R)\,i_r\quad R_s}{R_c+R+R_s}\ . \qquad (9)$$

The residual voltage is equal for $R_s < (R_C + R)$ during the nonconducting half-period in (9) to $R_s i_r$. At the same time the use of R_s reduces K_C; hence the value of R_s is chosen in such a manner as to arrive at a compromise between reducing the residual voltage and the value of K_C. Let us now derive expressions for K_C taking into consideration the value of R_s.

In the conducting half-period of the transistor we have

$$U'_{\text{cnd}} = \frac{E_c}{(R_c+R)/R_s + (R_c+R)/R_r + 1} - \frac{u_r}{R_r\ /R_s + R_r\ /(R_c+R)+1}. \qquad (10)$$

In the nonconducting half-period of the transistor we have

$$U'_{\text{non}} = \frac{R_s}{R_c+R+R_s}\,[E_c+(R_c+R)\,i_r] = \frac{R_s}{R_c+R+R_s}\,U_{\text{non}}. \qquad (11)$$

The voltage drop at the inverter output is

$$U'_{\text{out}} = \frac{R_s}{R_c+R+R_s}\left\{\frac{E_c+u_r}{R_r\ /R_s+R_r} \frac{[(R_c+R)/R_s+1]}{/(R_c+R)+1} + (R_c+R)\,i_r\right\}. \qquad (12)$$

When R_s is taken into account the conversion factor becomes

$$K'_C = \frac{R_s}{R_c+R+R_s}\left\{\frac{1}{R_r\ /(R_c+R)+1} + \frac{1}{E_c}\left[\frac{u_r\left(\dfrac{R_c+R}{R_s}+1\right)}{R_r\ /(R_c+R)+1} + (R_c+R)\,i_r\right]\right\}. \qquad (13)$$

For small values of $(u, i, R)_r$ it is possible to write (13) as

$$K'_C = \frac{R_s}{R_c+R+R_s}\cdot\frac{1}{R_r\ /(R_c+R)+1} = \frac{1}{\dfrac{R_c+R}{R_s}+1}\,K_C\ . \qquad (14)$$

Expression (14), which shows the relation of the inverter element parameters to K_C, can be considered as a basic formula for computing a stabilized inverter circuit for a low conversion frequency. The computation of an inverter by the above method differs from experimental data (for f_c = 50 cps) by no more than 3%.

The residual transistor parameters in the inverter circuit depend not only on temperature, but also on the amplitude and frequency of the converting voltage. For instance, if u_c is increased by 100% from 0.15 to 0.3 v the residual voltage u_r will rise from 1.5 to 3.9 mv, i. e., by 150%. The choice of the converting voltage value is made by means of the transistor output characteristics within a given range of measured dc voltages. For measuring thermocouple signals (0-15 mv) the converting (supply) voltage in inverters using transistors type P13, P14 and P16 must be 0.25-0.3 v [1].

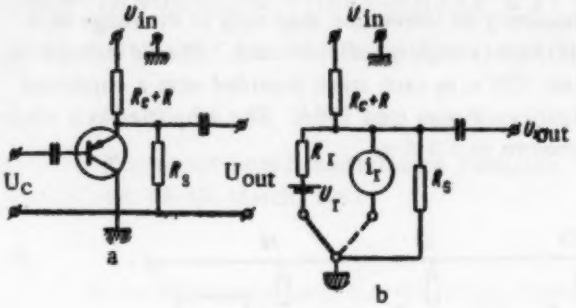

Fig. 2

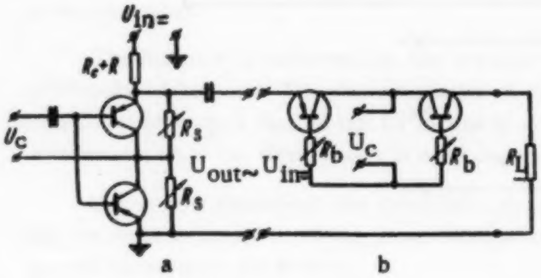

Fig. 3

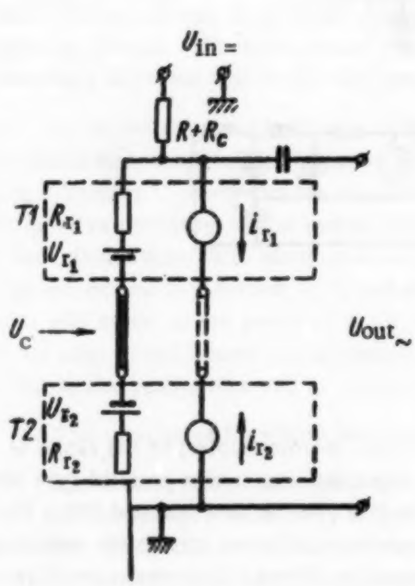

Fig. 4

The relation between the residual voltage at the output of the inverter and the frequency of the converting voltage is due to the existence of the p-n junction capacitance. If the inverter is fed with rectangular pulses the junction capacitance produces at the output, even for a $U_{in} = 0$, peaked signals in the form of differentiated leading edges of the conversion voltage. Since the capacitance of a blocked junction is larger than that of a conducting junction, blocking pulses produce peaks of a larger amplitude. If the charging time of the junction capacitance is small as compared with the conversion period, the peaks for $U_{in} = 0$ are very short. When the conversion voltage is raised the charging time of the junction capacitance becomes comparable to the conversion period and the distortion of the zero line increases considerably. This phenomenon limits the conversion frequency to the value of 0.01 f_{α}.

The sensitivity threshold of dc amplifiers can be lowered considerably by using balanced inverters with two transistors. The principle of operation of balanced inverters consists in connecting the two transistor converting circuits in parallel and their measuring circuits in series opposing. In such an inverter the transistor residual parameters are partly or completely (if they are equal) balanced out. Two basic circuits for balanced inverters of a parallel and series type are shown in Fig. 3,a and b. Figure 4 shows an equivalent circuit of a balanced parallel inverter.

It will be seen from the equivalent circuit that the residual currents and voltages cancel in a balanced connection, but the residual resistances add. Therefore, a balanced inverter has smaller values of u_r and i_r but a larger value of R_r than an ordinary inverter. Since the mass-produced transistors have a large dispersion in their residual parameters, it is desirable for a full compensation of a balanced inverter to select transistors with the same values of u_r and i_r or employ circuit compensation. Circuit compensation of residual parameters in a balanced inverter can be attained by connecting resistor R_s in parallel with the transistor. Experiments have shown that the optimum values for such resistors in parallel inverters using transistors type P13, P14 and P16 lie in the range of 300-1000 ohm. With an individual selection of transistors of the above types a balanced inverter can be used for measuring dc voltages from 0.1 mv upwards (with a ratio of the effective to the residual voltage of the order of 2.5). When compensating resistors are used a rise in the ambient temperature up to 50°C does not produce a noticeable increase in the residual voltage at the output of the inverters. If such resistances are not used, the temperature compensation of balanced inverters will only be reliable for transistors with extremely close values of residual parameters. In certain transistors with equal residual parameters at room temperature a difference of 20% may arise in these parameters at a temperature of 50°C.

Figure 5 shows a schematic of a dc amplifier for the range of 0-50 mv with a balanced inverter of a parallel type using P13 transistors (T1 and T2). The basic error of the amplifier is about 1%. Additional errors due to variations in the nominal supply voltage of -15 to +10% and in the ambient temperature from 15 to 50°C do not exceed 2.5%. The residual voltage referred to the input amounts to less than 0.3 mv, the total drift of the amplifier referred to the

input is 0.3 mv (for 8 hr). The source of the converting voltage consists of a balanced multivibrator using transistors type P13 (T3, T4) with an additional matching stage (T5). The frequency of conversion may vary in the range of 1-4 kc, thus making it possible to use the amplifier for signals with relatively high speed variations. The ac voltage is amplified in a four stage amplifier using transistors P14 (T6-T8 and T9) with each stage provided with a combined series-parallel negative feedback. The output rectifier consists of silicon diodes type 206A. The amplifier as a whole has a negative feedback with a factor of 0.2. Its output voltage amounts to 0.3-5 v.

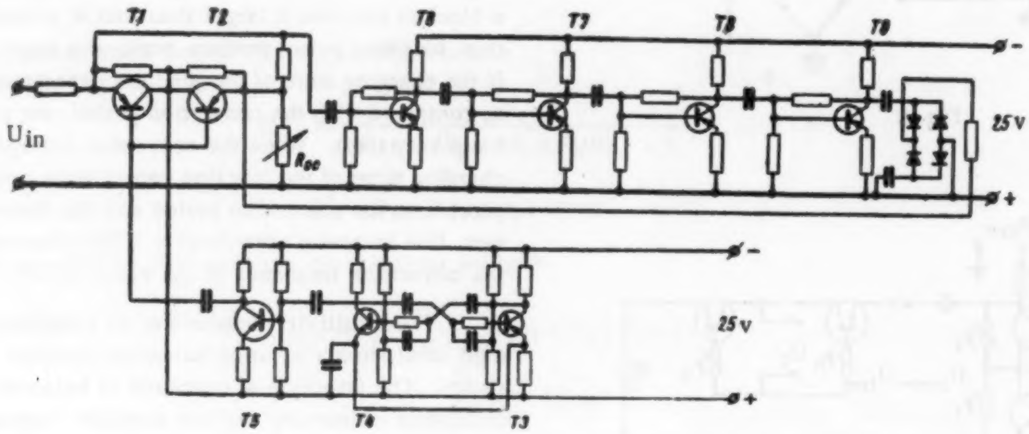

Fig. 5

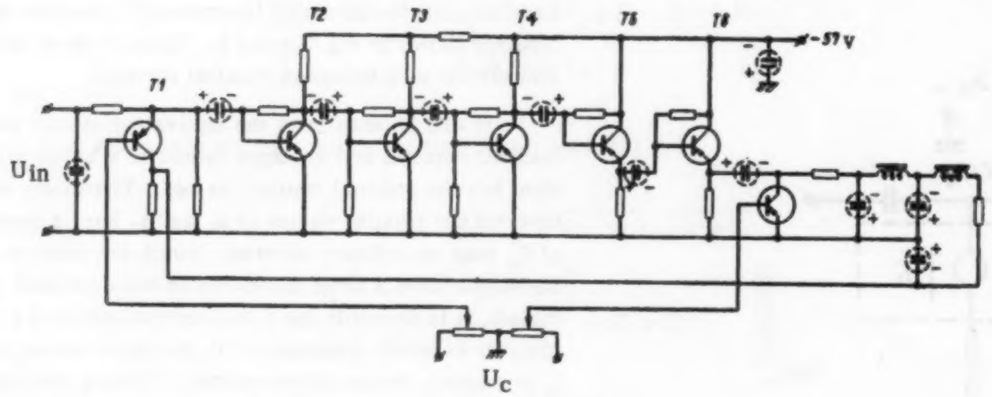

Fig. 6

Figure 6 shows a circuit for an amplifier of slowly changing dc signals (from thermocouples) in the range of 0-15 mv. The mains 50 cps voltage is used here for conversion purposes. The input inverter is of a parallel type with a P16 transistor (T1). The ac amplifier has five stages and uses transistors types P16 (T2-T4 and T5 and T6). The preamplifying stages have a direct parallel feedback; the output stages use a common collector circuit for matching to the output integrator. The total drain of the amplifier from the source amounts to 50 ma. Its output current into a load of 1 kilohm is 5 ma. The ac amplifier is connected to a transistor phase-sensitive converter (T7) and an integrator. The basic error of the amplifier is 0.5%. The additional errors due to variations of the supply voltage from -15 to +10%, of the ambient temperature in the range of 10-50°C, of the converting voltage from -15 to +20%, and of the load in the range of 800-1200 ohm, do not exceed 1.5%. Its feedback factor is 0.05. The laboratory model of this amplifier is mounted in a cabinet measuring 15 × 20 × 20 cm. The total amplifier drift referred to its input does not exceed 0.1 mv during 8 hr.

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LOW-INDUCTION THREE-PHASE WATTMETER FOR COMMERCIAL AND HIGHER FREQUENCIFS

V. S. Popov and Yu. A. Mal'tsev

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In the Electrochemical Institute of the USSR Academy of Sciences the authors developed a multirange low-induction wattmeter for the range of 50-2000 cps for measuring power by the equal temperature method [1].

In the above method of measurement the thermal effect which is proportional to the measured power is compared with a similar thermal effect produced by a direct current. The equality of powers is gauged by the equality of temperatures.

The balance is indicated by two metallic heated resistors (R_1 and R_2) placed in air, each having three electrically insulated heaters. The characteristics of the heater resistors are similar but can be of an arbitrary type. The measuring process is carried out by means of three heaters of the second heated resistor (R_2) and two heaters of the first resistor (R_1); the third heater is only required for constructional symmetry and serves as a spare.

Let us now determine the conditions under which the heated resistor with several electrically insulated heaters will be equally sensitive to the power dissipated in any of them, including the power dissipated in the corresponding control resistors of the heaters.

Let us assume that in a heated resistor with several heaters the ohm-ampere characteristics $R = F(I)$ have a difference $\underline{n}$ expressed in terms of the resistance increment above its value at a nominal heating current. The above value can be reduced to $\underline{m}$ if the heaters are shunted by resistors and it is assumed that current I is the total current before it divides. The resistances of the two branches, as a rule, are not equal, and therefore the heated resistance possesses a different sensitivity with respect to the voltage at the branching.

If the building-out resistance is connected in series with the two branches and the heater circuit resistances are equalized, the relative deviation of the volt-ohm characteristics of heated resistors $R = F(U)$ will also be equal to $\underline{m}$. Voltage U represents the total voltage across the building-out resistor and the branching. It is obvious that the relative deviation of the heater resistance characteristic $R = F(P)$, where $P = UI$, will in this case also be equal to $\underline{m}$. In fact, if the maximum absolute deviation of the ohm-ampere characteristics of the heater resistances is equal to l ohm and occurs at a current $I = I_1$ and a voltage $U = U_1$, the maximum absolute deviation of the ohm-watt characteristics will occur at the power of $P_1 = U_1 T_1$ and will also be equal to l ohm. Considering that it is possible to provide an equality of the heater circuit resistances with high precision, the relative deviation in the ohm-watt characteristics of the heater resistances will in practice be equal to the relative deviation of their ohm-ampere characteristics.

The conditions for which two different heater resistors have equal ohm-watt characteristics are wholly identical.

The heater resistances differ considerably from each other in their value; therefore the difference in their ohm-watt characteristics is relatively large. However, the temperature of the heater resistances has the same relation to the power dissipated in their heater circuits. The deviation of these characteristics is equivalent to that of ohm-watt or ohm-ampere characteristics of one heated resistor with several heaters.

In heated resistors of the IÉM type deviations in ohm-ampere characteristics with a shunt do not exceed m = 0.02-0.03%, and without a shunt n = 0.30-0.35%. For relatively rough measurements heaters need not be shunted, but heater building-out resistances remain necessary.

The wattmeter we have developed consists of two single-phase converters and a measuring bridge with automatic balancing. The converters are connected in a two-wattmeter circuit (Fig. 1). Each converter consists of a voltage transformer, a shunt and two heaters.

By means of adjustable building-out resistors in the heaters (not shown in Fig. 1) the following condition can be met:

$$R_{LT} = R_{ST} = R_{HE} = R_{EP} = R.$$

Through heater 1 of the first converter a current will flow equal to

$$i_1=\frac{1}{R}\left(\frac{u_{ab}}{2k}+i_a\frac{R_S R}{2R_S+R}\right),$$

and through heater 3 a current equal to

$$i_3=\frac{1}{R}\left(\frac{u_{ab}}{2k}-i_a\frac{R_S R}{2R_S+R}\right).$$

In the second converter the current flowing through heater 2 will similarly be equal to

$$i_1'=\frac{1}{R}\left(\frac{u_{cb}}{2k}+i_c\frac{R_S R}{2R_S+R}\right),$$

and through heater 4 it will be equal to

$$i_2'=\frac{1}{R}\left(\frac{u_{cb}}{2k}-i_c\frac{R_S R}{2R_S+R}\right),$$

where u_{ab}, u_{cb}, i_a and i_c are the linear voltages and currents; $\underline{k}$ is the transformer ratio; R_S is the resistance of the shunt.

The mean powers dissipated in resistances R_{LT}, R_{HE}, R_{ST} and R_{FE} amount respectively to

$$P_1=\frac{1}{TR}\int_0^T\left(\frac{u_{ab}}{2k}+i_a\frac{R_S R}{2R_S+R}\right)^2 dt;$$

$$P_2=\frac{1}{TR}\int_0^T\left(\frac{u_{cb}}{2k}+i_c\frac{R_S R}{2R_S+R}\right)^2 dt;$$

$$P_3=\frac{1}{TR}\int_0^T\left(\frac{u_{ab}}{2k}-i_a\frac{R_S R}{2R_S+R}\right)^2 dt;$$

$$P_4=\frac{1}{TR}\int_0^T\left(\frac{u_{cb}}{2k}-i_c\frac{R_S R}{2R_S+R}\right)^2 dt,$$

where T is the period of the alternating current.

The difference of powers $(P_1+P_2)-(P_3+P_4)$ is directly proportional to the mean value of the measured power P_x:

$$(P_1+P_2)-(P_3+P_4)=\frac{2R_S}{kT(2R_S+R)}\int_0^T(u_{ab}i_a+u_{cb}i_c)\,dt=$$

$$=\frac{2R_S\,P_x}{k(2R_S+R)}=k_1P_x, \tag{1}$$

where

$$k_1=\frac{2R_S}{k(2R_S+R)}.$$

The above expression does not depend on the nature of the heater resistances, hence not on their construction, their properties or the composition of the surrounding medium, etc.

The difference of powers $(P_1+P_2)-(P_3+P_4)$ can be measured by a balancing method. For this purpose heater resistor R_2, whose temperature during measurements is lower than that of resistor R_1, is provided with an additional heater 5. For a certain value of the current flowing through the additional heater the temperature of both heater resistors will be equal; at the same time their total powers dissipated in the heater and additional resistors of the first and second heated resistances will be equal. Hence, $P_{ad}=k_1P_x$, where P_{ad} is the power dissipated in heater 5 and in its building-out adjustable resistor.

Hence voltage U_{ad} across heater 5 and its building-out resistor will be equal to

$$U_{ad} = \sqrt{P_{ad}(R_h + R')} = \sqrt{k_1 P_x (R_h + R')},$$

where R_h is the heater resistance; R' is the building-out resistor.

Since the heater circuit resistances in resistors R_1 and R_2 must be equal, the value of R' will amount to

$$R' = R - R_h.$$

Hence, the voltage drop across the additional heater will be equal to

$$U'_{ad} = \frac{R_h}{R}\sqrt{k_1 R P_x} = c\sqrt{P_x}, \qquad (2)$$

where c is a constant.

It will be seen from (2) that the voltage across the heater is directly proportional to the square root of the measured power.

The coefficient of proportionality c of the wattmeter can be taken care of in calibration. Thus, it is not necessary to have a building-out resistance for heater 5.

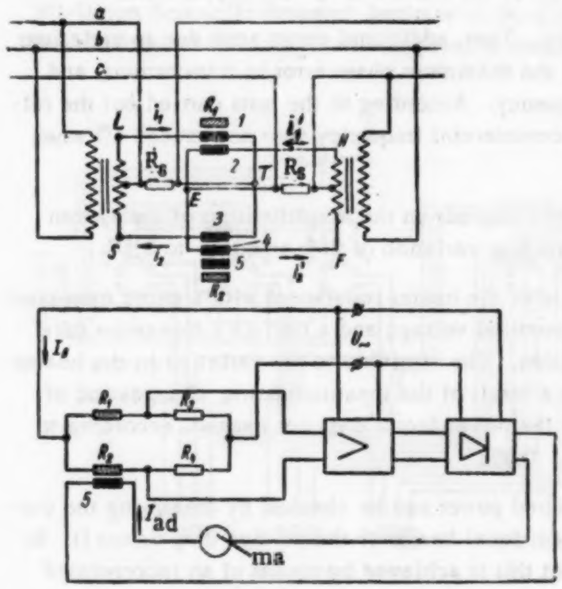

Fig. 1

Similar results can be obtained if current, instead of voltage, is taken as the output parameter of the wattmeter. The possibility of dispensing with the building-out resistor in this case is obvious.

The heater resistors R_1 and R_2 are connected into two adjacent arms of the measuring bridge, which is fed from an auxiliary source of alternating current. The additional heater 5 is connected into the bridge diagonal through an amplifier. The measured power unbalances the bridge. The unbalanced voltage amplified by the amplifier feeds the additional heater and tends to balance the bridge. In a steady state condition the bridge imbalance will be determined by the statics in the system, which can be made sufficiently small.

Let us now examine the requirements for separate elements of the wattmeter.

<u>Heater resistors</u>, developed at the Electromechanical Institute of the USSR Academy of Sciences, consist of a platinum helix wound over three nichrome heaters placed side by side. Each heater is covered with a glass insulating tube.

The time constant of the heater resistors used in the instrument amounts to 0.8-0.9 sec.

The diameter of a nichrome wire is 20 μ. An increase in the size of the diameter raises the inertia of the heater resistor, whereas its decrease worsens the vibration- and shockproof properties of the resistor. The diameter of the platinum wire is 18μ. The thickness of the glass insulating tube does not exceed 15 μ. The break-through voltage between the heaters amounts to 320-350 v (U_{max} = 450 − 500 v). The heater resistance is 45-50 ohm. Helix resistance at t = 20°C is approximately 20 ohm. The heater resistors are placed in pairs in a nonvacuum brass bulb, which serves to protect them from mechanical damage.

<u>Voltage transformers</u> are designed for the primary voltage of 220 v and a secondary voltage of 6 v. The secondary voltage is provided with a center tap for connection to the shunts.

With a rising frequency the phase difference between the measured voltage and current in the heater increases owing to the phase error in the transformers. In order to compensate for this error building-out resistor r is connected in series with each heater, and shunted by a capacitance (Fig. 2). As a result of this the angular error of the wattmeter has been reduced to 0.0015 rad in the range of 50-2000 cps.

Shunts. The wattmeter has four current measuring ranges (5, 1, 0.5 and 0.25 amp), and four pairs of shunts for them. For a nominal current the voltage drop across each shunt amounts to 0.8 v. The shunts are switched by means of a special switch whose contact resistance is small and stable.

Amplifier and phase-sensitive rectifier. The amplifier has three stages. The first two stages consist of 6N2P tubes, and the third of a 6P1P tube. In order to raise its gain stability the output stage has an internal feedback.

The phase-sensitive rectifier consists of a controlled transistor P-4 and two rectifiers type DGTs. One of the rectifiers is connected in series with the collector of the transistor and the other shunts the output transformer winding during the half-period when the controlled transistor is blocked. Control of the transistor, and supplies to the bridge and the amplifier, are obtained from the same transformer.

Let us now examine the wattmeter errors. An analysis of the operation of a bridge with heater resistors shows that the relative errors due to variations in the bridge supply voltage, in the gain of the amplifier and in its output impedance are inversely proportional to the system's amplification factor K [2]. For large values of K oscillations arise in the system. The wattmeter has a $K \approx 30$ and its errors do not exceed 0.3-0.6%.

During measurements the heater resistors are balanced, i. e., their temperatures are equal. Variations in the ambient temperature affect in the same way both heater resistors without unbalancing the bridge or influencing the measurement results.

According to the tests carried out by us the temperature error of the converter is negative and amounts to 0.4% for temperature variations of 10°C. The temperature error is due to variations in the resistances of heaters, transformer windings and rectifiers. It can be easily compensated by winding part of the resistance R_4 with copper wire.

At higher frequencies the transformer phase errors begin to vary. Thus, additional errors arise due to variations in the power factor. In wattmeters covering a range of 50-2000 cps the maximum phase error in transformers, and hence the maximum frequency error, occur at the commercial frequency. According to the tests carried out the relative referred wattmeter error due to the power factor effect at the commercial frequency does not exceed 1% when nominally $\cos\ \varphi = 0.15$.

The error due to voltage variation in the measured power circuit depends on the amplification of the system and drops with the rise of gain. The maximum referred error for a voltage variation of 20% amounts to 0.8%.

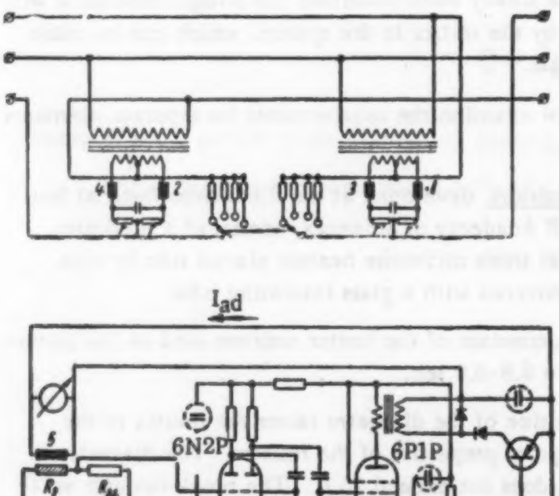

Fig. 2

The rise in the heater resistances with a rising measured power for a nominal voltage and a $\cos\ \varphi = 1.0$ is taken care of in calibration. The error due to the variation in the heater resistance as a result of the measured power changes and of variations in the power factor does not exceed, according to calculations, 0.3%.

The measured power can be checked by measuring the current in the additional heater or the voltage drop across it. In our instrument this is achieved by means of an incorporated grade 1.0 millivoltmeter, which is connected in parallel with the heater. The front panel of the set is provided with terminals for connecting a more accurate instrument to measure the voltage drop across the heater. The wattmeter is calibrated at the commercial frequency.

Main technical characteristics of the wattmeter. Accuracy grade 1.5 (with a grade 0.5 millivoltmeter); damping time 4.0 sec; effective frequency range 50-2000 cps; nominal power factors 1.00-0.50-0.15; nominal voltage 220 v; nominal currents 5.0-1.0-0.5-0.25 amp; nominal voltage drop across the shunts 0.8 v; the current in the wattmeter parallel circuits does not exceed 2 ma.

The wattmeter consists of a compensating (balanced) system and can be used as a transmitting device for remote measurement of power.

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ATTACHMENT TO THREE-PHASE INSTALLATIONS

V. S. Sheiko and P. I. Vikulin

Translated from Izmeritel'naya Tekhnika, No. 3, p. 44, March, 1961

The above attachment consists of a commutating device by means of which it is possible rapidly to obtain any wattmeter connecting system for checking electricity meters. Three-phase installations made by the "Étalon" and "Vibrator" Plants, as well as installations of earlier types, have no special switch for connecting wattmeters for different test procedures without reconnecting the conductors on the instrument terminals. At the same time the experience of working with three-phase installations types PTU-1 and PTU-2 made by the Repair and Experimental Shop of the All-Union Scientific Research Institute of the Committee of Standards, Measures, and Measuring Instruments has shown the great advantages of using such a switch.

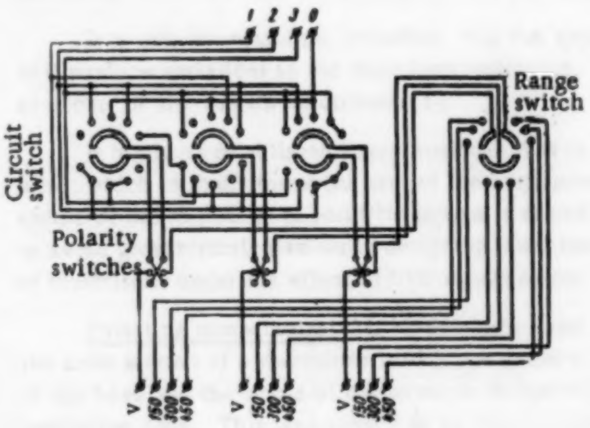

Fig. 1.

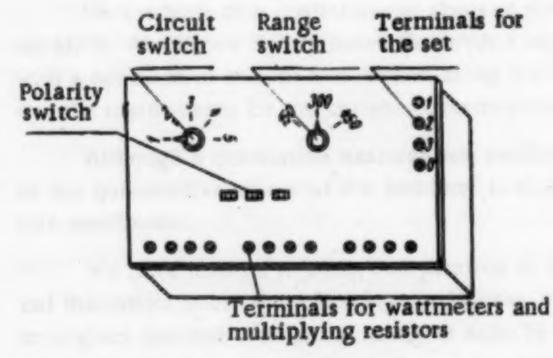

Fig. 2.

A wattmeter circuit switch makes the work of state inspectors considerably easier, eliminates errors in assembling circuits, and saves time which is normally required for the assembly.

On the authors' suggestion the Committee's Research Institute has developed and made a device for switching wattmeter circuits (Fig. 1) made in the form of a separate attachment measuring 140 × 230 × 50 mm, and weighing 1 kg, which can be easily used with any three-phase installation by simply connecting it to the set terminals. The wattmeters and multiplying resistors are connected to the appropriate terminals on the attachment (Fig. 2).

For ease in operation, in order to eliminate completely any switching on the wattmeters themselves, the attachments are also supplied with wattmeter range and polarity switches.

By means of this attachment it is possible to obtain five principal wattmeter connection circuits (Fig. 3):

a) a circuit for measuring power by means of two wattmeters in phases 1-2;

b) same, but with wattmeters connected to phases 1-3;

c) a circuit used in checking four-conductor electricity power meters connected to phases 1, 2, 3;

d) a circuit used for checking wattless power electricity meters with three series windings connected to phases 1, 2, 3;

e) a circuit for measuring wattless power by means of two wattmeters in checking reactive electricity meters with a 60° phase shift.

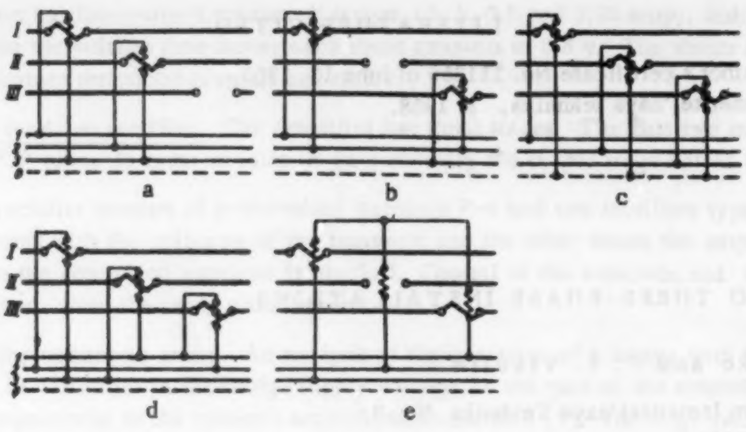

Fig. 3.

The above attachment should find a sphere of application both in state inspection laboratories for measurement equipment and in electricity meter repair workshops of the service inspection area laboratories and other organizations which have three-phase installations. It is therefore advisable to organize mass production of these attachments.

HIGH AND ULTRAHIGH FREQUENCY MEASUREMENTS

WORKING CHARACTERISTICS OF THERMISTORS SUBJECTED TO UHF SIGNAL PULSES

V. D. Frumkin

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 45-51, March, 1961

The application of thermistors for measuring the mean power of UHF signal pulses is based on the conception of a thermistor as an inertial element whose resistance does not depend on the instantaneous power value of relatively rapid processes, but is only determined by the mean power of the signal. Moreover, it is assumed that the thermistor does not react to the electric field of a relatively high intensity developed in it during the action of the pulse. These assumptions are the basis for measuring the mean power of a UHF signal pulse without any limitations regarding the parameters of the signal envelope, and without consideration of additional errors due to operation conditions.

The only instrument known to us with a specified limitation is the widely used type VIM-1 power meter. The specifications for this instrument provide a maximum pulse power of 6 w for its thermistor heads. However, the study of this instrument has shown that the above limitation is not based on the characteristics of the thermistor operation or those of the bridge, and in some instances does not protect the instrument and in others unreasonably limits its application.

A study of the behavior of thermistors fed by a current of several kc has shown that they behave under these conditions as elements possessing a small inertia due to which they distort the waveform of the voltage across the thermistor when fed with a purely sinusoidal current, produce a phase-shift between the voltage and the current, etc.

It is natural to assume, therefore, that the application of a UHF signal of varying intensity to the thermistor will produce variations in the thermistor resistance, which may in turn lead to measurement errors in the same manner as occurs in low-inertia bolometers [1].

It has been established experimentally that in measuring the mean power of a UHF signal pulse additional errors arise, which in many instances exceed the basic error of the instrument [2]. This is confirmed by the fact that the response of thermistors to pulsed UHF signals is essentially different from their response to unmodulated signals. In order to avoid gross miscalculations in designing thermistor measuring devices it is absolutely essential to know the behavior of thermistors under the effect of UHF signal pulses.

Pulsating temperature. M. V. Abrosimov and L. A. Lyubimov were the first to note that the current density along the cross section of a thermistor lead is not uniform. In fact the current flows only through an extremely small part of the bead and the whole of the power is dissipated in that small part, which M. V. Abrosimov calls the power concentration zone. This idea proved to be fruitful and provided a satisfactory explanation of the behavior of a thermistor fed by audio-frequency currents. L. A. Lyubimov explains the current concentration by the thermistor configuration and its nonuniform structure.

On the basis of a qualitative analysis of the process of current stabilization along the cross section of a thermistor M. V. Abrosimov rightly noted that with a negative resistance temperature coefficient a balance will be attained with a nonuniform current distribution along the cross section of the thermistor, and came to the conclusion that this was the main reason for the current concentration.

Although a qualitative examination confirms the existence of current concentration, it does not throw any light on the quantitative aspect of the problem, i. e.,on the degree of current concentration due to the negative temperature coefficient.

We have made a rigorous computation of the nonuniform current density along the cross section of a cylindrical thermistor under the following conditions: a) the thermistor length is sufficiently greater than its radius to be able to neglect thermal conduction along its axis; b) the thermistor is fed from a current source and hence the power dissipated in the thermistor is limited, and the temperature in any part of its cross section cannot attain infinity; c) the conductance temperature coefficient does not depend on temperature.

The latter assumption does not contradict the physical meaning of the problem and can only lead to an exaggerated evaluation of the degree of current concentration.

With the above assumptions the thermal equilibrium equation and the limiting conditions will be written in the form

$$\frac{d^2T}{dr^2}+\frac{1}{r}\frac{dT}{dr}+\frac{1}{\varkappa}F=0; \tag{1}$$

$$\left(\frac{dT}{dr}+hT\right)_{r=a}=T_m, \tag{2}$$

where T is the temperature; $\varkappa$ is the thermal conductivity factor; $\underline{a}$ is the thermistor radius; $\underline{h}$ is the heat exchange factor; T_m is the ambient temperature; $F=\sigma E^2$ is the density of thermal sources; σ is the conductance; E is the electrical field.

The solution of (1) which does not become infinite at any point of the cross section can be represented, taking condition (2) into consideration, by the formula

$$T=T_m+\frac{1}{\alpha}\left[\frac{I_0(pr)}{I_0(pa)-\frac{p}{h}I_1(pa)}-1\right], \tag{3}$$

where α is the conductance temperature coefficient;

$$p=E\sqrt{\frac{\sigma_0\alpha}{\varkappa}};$$

σ_0 is the thermistor conductance at ambient temperature; I_0 and I_1 are Bessel functions of the zero and first order respectively.

The maximum ratio of current densities along the cross section of a thermistor with parameters similar to those of type TSh-2 thermistor was calculated from (3).

This ratio proved to be smaller than (1), i. e., the current concentration due to a negative temperature coefficient proved to be insignificant.

Current concentration depends to a much greater extent on the configuration of the thermistor. Thus, for instance, a conducting sphere with lead-out conductors at the poles has a ratio of current densities at the center and at the periphery of an equatorial cross section greater than 2, without taking into consideration the effect of a negative temperature coefficient.

If it is taken into account that in an actual thermistor bead the lead-out conductors are taken to a considerable depth inside the sphere, it becomes clear that the configuration characteristics of the thermistor constitute the main reason for the current concentration.

In our further analysis of the thermistor operation in a pulsed condition we shall assume that practically the whole of the power fed to the thermistor by the initial heating current and by the ultrahigh frequency current is concentrated in a relatively small portion of the thermistor, which we shall call the active zone, and that the remaining mass serves only for the conduction of heat. Conclusions arrived at on the basis of such a theoretical analysis agree well with the experimental data.

Let us first examine the heat conduction process of short periodic heat pulses produced by a point source in an infinite homogeneous isotropic space which possesses the parameters of a semiconductor. The solution of such a problem provides a good idea of the nature of the heat exchange process inside a semiconductor bead and provides correct premises for the solution of subsequent problems.

The thermal equilibrium equation for the above case has the form

$$\frac{\partial T}{\partial t}=\frac{\varkappa}{c\rho}\Delta T+\frac{F}{c\rho}, \tag{4}$$

where t is the time; c is the specific heat of the semiconductor; ρ is semiconductor density; and Δ is the Laplace operator.

If a point source is used, $F=\infty$ at the point where the source is placed, and $F=0$ at all other points in space.

For points outside the source, (4) assumes the form

$$\Delta T=\frac{c\rho}{\varkappa}\cdot\frac{\partial T}{\partial t}. \tag{5}$$

In order to evaluate the source let us surround it with a sphere of radius r and calculate the total heat flow passing through the sphere per unit of time. It is obvious that for $r\to 0$ this flow will be equal to the power of the source, i. e.,

$$\lim_{r\to 0}\oint \bar{g}\,d\bar{s}=P(t), \tag{6}$$

where $\bar{g}=-\varkappa\Delta T$ is the flow through a unit surface; $d\bar{s}$ is a surface element; and $P(t)$ is the power of the source.

In this formula and subsequently the measured pulsed power is considered as the power of the source. The heating power produces a stationary heat flow which is of no interest to us in this analysis.

Function $P(t)$ is expressed in the following manner: for the duration of the pulse ($nt_c \le t \le nt_c + t_p$)

$$P(t)=P; \tag{7a}$$

for the interval between pulses

$$(nt_c+t_p<t<[n+1]\,t_c),\ P(t)=0, \tag{7b}$$

where t_p is the duration of the pulse; t_c is the duration of the whole period; and $n=0,1,2,3 \;\ldots$.

Considering that the space is homogeneous and isotropic, and that we are dealing with a symmetrical sphere, it is possible to transform (5) and condition (6) to the form

$$\frac{\partial^2 T}{\partial r^2}+\frac{2}{r}\,\frac{\partial T}{\partial r}=\frac{c\rho}{\varkappa}\,\frac{\partial T}{\partial t}, \tag{8}$$

$$\lim_{r\to 0}-4\pi r^2\varkappa\frac{\partial T}{\partial r}=P(t). \tag{9}$$

By solving (8) with the help of (9) we obtain

$$T(r,t)=\frac{P}{4\pi^2\varkappa r}\sum_n\frac{1}{n}\left[\sin n\Omega t_p-j\,(1-\cos n\Omega t_p)\right]e^{-\alpha_n r}\cdot e^{j(\Omega nt-\alpha_n r)}, \tag{10}$$

where Ω is the angular pulse repetition frequency; and $\alpha_n=\sqrt{\Omega n c\rho/2\varkappa}$ is the propagation constant.

Expression (10) shows that heat propagation from a point source has the characteristic of a wave. The propagation constant indicates both the phase shift of the temperature wave and its attenuation expressed in nepers. Since α_n rises proportionately to $\sqrt{n}$, the attenuation increases for higher harmonics.

At a frequency of 50 cps $\alpha_n = 160$ nepers/cm for the semiconductor parameters of $\varkappa = 2\cdot 10^{-2}$ w/deg·cm; $c=0.8$ joule/g·deg; and $\rho=4$ g/cm^3.

If the radius of a thermistor bead is assumed to be 0.15 mm, the temperature wave on reaching the surface of the bead will be attenuated approximately by a factor of 11.

It is obvious that in such a case the conditions at the surface of the bead will have practically no effect on the heat exchange between the region where the power is produced and the remaining mass of the semiconductor.

For the above solution it is possible to arrive at an important conclusion, that in the case of a concentrated power being periodically dissipated by short pulses in a relatively small portion of a semiconductor bead, the heat exchange for a duration of the order of a pulse repetition period can be considered as a heat exchange between the

active region of the thermistor and its remaining mass, without taking into account the conditions at the surface of the bead. Naturally, the time constant of such a heat exchange will be considerably smaller than that of a bead as a whole with the surrounding medium.

The solution of this problem by assuming that the heat is produced at one point obviously does not provide any information on the law of temperature changes in the active zone of the thermistor. This information, however, is indispensable since we assumed that the entire direct current passes through this zone and since we are interested in the thermistor resistance variations due to this current.

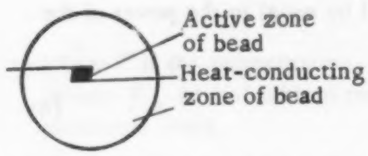

Fig. 1. Idealized physical model of a semiconductor bead.

In order to obtain this information it is necessary to solve the problem of the heat exchange between the active zone and the surrounding mass of the thermistor, taking into account the finite dimensions of the zone.

We shall make our computations for an idealized physical model (Fig. 1) of the bead with respect to which we shall make the following simplifying assumptions: the entire power transmitted to the thermistor by the UHF pulse signal is dissipated in the active zone uniformly with respect to its volume; the temperature of the whole active zone at each instant is the same; the specific heat of the zone and its heat-transfer coefficient to the surrounding mass of the thermistor do not depend on temperature; the heat dissipation through the current conductors is negligibly small; the boundary conditions at the surface of the bead do not affect the nature of the heat exchange, i. e., the heat dissipation is the same as it would be in an infinite space possessing the parameters of a semiconductor.

The heat equilibrium equation will then be

$$C\frac{d\Theta}{dt}+K\Theta=P(t), \tag{11}$$

where C is the thermal capacity of the zone; K is the heat-transfer coefficient; Θ is the temperature variation; P(t) is the measured power as given in (7a and 7b). The solution of (11) for a steady state condition with $t \to \infty$ has the form:

for the duration of a pulse

$$(nt_c<t<nt_c+t_p)$$

$$\Theta_1=\Theta_0\left[1-e^{-\frac{t}{\tau}}\frac{1-e^{-\frac{t_c-t_p}{\tau}}}{1-e^{-\frac{t_c}{\tau}}}\right]; \tag{12}$$

for an interval between pulses

$$(nt_c+t_p<t<[n+1]\,t_c)$$

$$\Theta_2=\Theta_0\frac{1-e^{-\frac{t_p}{\tau}}}{1-e^{-\frac{t_c}{\tau}}}\cdot e^{-\frac{t-t_p}{\tau}}. \tag{13}$$

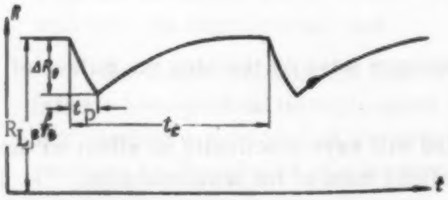

Fig. 2. Relation between the thermistor resistance and time, calculated from (21).

In (12) and (13) Θ_0 = P/K is the temperature increment which would have existed in the concentration zone if power P were dissipated continuously, and τ=C/K is the heat-exchange time constant.

Tests have shown that for various thermistors τ is of the order of several tens or hundreds of microseconds. Since the duration of the measured pulses does not as a rule exceed 10 msec, and the pulse repetition period is of the order of a few units or tenths of milliseconds, the following relation holds in the majority of practical cases:

$$t_p \leqslant 0.1\tau; \quad \tau<0.1t_c. \tag{14}$$

and with an error of the second order of magnitude (12) and (13) can be expressed in the form

$$\theta_1 = \frac{P}{C} t ; \tag{15}$$

$$\theta_2 = \theta_0 \frac{t_p}{\tau} e^{-\frac{t-t_p}{\tau}} = \frac{P t_p}{C} e^{-\frac{t-t_p}{\tau}} . \tag{16}$$

The temperature increment during the operation of the pulse is

$$\theta_p = \frac{P t_p}{C} = \frac{W_p}{C} , \tag{17}$$

where W_p is the energy contained in one pulse.

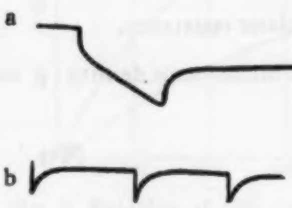

Fig. 3. Oscillograms of a thermistor dc resistance pulsation under the effect of a pulsed signal, t_p = 5μ sec and F = 450 pps. a) During pulsations; b) in the interval between pulses.

Expressions (15) and (16) have a simple physical meaning. In fact, for such a short time interval as the duration of a pulse as compared with the time constant, only an insignificant fraction of the energy dissipated in the active zone will be transmitted to the surrounding medium. In such a case the rise of temperature for a constant power will follow a linear law. The rate of temperature rise is inversely proportional to the specific heat of the heated object. For the duration of the pulse the temperature increment is equal to the energy dissipated in that time divided by the thermal capacity (17). Formula (16) represents the fact that the active zone, having reached in time t_p a temperature of Θ_p, cools in the intervals between pulses according to an exponential law with a time constant of τ.

The absolute temperature of a thermistor active zone heated by dc pulsed signals can be expressed as a sum of two components, one of which does and the other does not depend on time:

$$T_1(t) = T_0 + \frac{\Theta_p}{t_p} t \quad (n t_c < t < n t_c + t_c); \tag{18}$$

$$T_2(t) = T_0 + \Theta_p e^{-\frac{t-t_p}{\tau}} \quad (n t_c + t_p < t < [n+1] t_c). \tag{19}$$

In these expressions

$$T_0 = \frac{P_0}{A} + \frac{P t_p}{t_c} \left(\frac{1}{A} - \frac{\tau}{C} \right) + T_m , \tag{20}$$

where P_0 is the thermistor dc heating power; and A is the heat-transfer coefficient of the thermistor as a whole to the surrounding medium.

Since in practice all the current which heats the thermistor is concentrated in the active zone, the dc thermistor resistance can be represented by the expression

$$R(t) = R_L e^{\frac{\beta}{T(t)}} , \tag{21}$$

where R_L and β are the thermistor parameters, and T(t) is determined from (18) and (19).

The relationship of R(t) is shown in Figure 2.

It will be seen from the oscillogram (Fig. 3) that the calculated and experimental laws of the thermistor resistance pulsations are in good agreement. The difference between the two during the pulse operation consists, in addition to a gradual decrease of the resistance, in an effect which coincides with the leading edge of the pulse and disappears at the end of the pulse. This effect will be explained below.

The completion of the cooling process in a time which in practice is shorter than the pulse repetition period indicates the small time constant of the process.

The amplitude of resistance pulsations and the time constant for different beads and different types of thermistors may differ considerably with the same signal parameters.

For cylindrical thermistors, for instance of the TSh-2 type, the amplitude of resistance pulsations will amount, with the remaining conditions unchanged, to a few tenths of that produced in bead thermistors, thus indicating a considerably larger active zone volume in thermistors of a cylindrical type. Such essential differences are due to the configuration of the thermistor volume and the construction of its lead-out conductors.

The maximum variation of the thermistor resistance in a general case is represented by expression

$$\Delta R_\theta = R(0) - R(t_p) = R(t_p) - R(t_c). \tag{22}$$

However, for small variations of temperature it is possible to assume the maximum resistance variation to be proportional to the maximum temperature change, i. e.,

$$\Delta R_\theta = \alpha_\theta \, \Theta_p R_0. \tag{23}$$

where α_θ is the relative temperature sensitivity of the thermistor, and R_θ is the dc thermistor resistance.

If the thermal capacity of the zone is expressed by means of its volume V_z, the semiconductor density ρ and its specific heat c, it is possible to obtain from (17) the expression

$$\Delta R_\theta = \frac{R_0 \alpha_\theta P t_p}{V_z c \rho}. \tag{24}$$

With the knowledge of the semiconductor and signal parameters it is easy to determine V_z if ΔR_θ is measured.

ΔR_θ was measured on an oscillograph which was calibrated for amplitude deviations of resistance pulsations in a thermistor connected to the arm of a dc-fed bridge.

The volume of the active zone was evaluated in 8 thermistors type T9 as $0.5 \cdot 10^{-6}$ to $0.08 \cdot 10^{-6}$ cm^3, which for a bead diameter of 300 μ amounted to 1/30 to 1/200 of the bead volume. The relatively small volume of the zone thus obtained justified the assumptions made at the beginning of this computation.

Effect of the high-frequency field strength on the thermistor. Variations in conductivity. In the majority of bead-type thermistors one observes during the operation of a pulse, in addition to a gradual reduction of resistance due to a rising temperature in the active zone, also an instantaneous variation of resistance practically coinciding with the leading edge of the pulse (Fig. 3), a variation which cannot be due to heat processes.

In order to explain this phenomenon it is necessary to examine the theory of semiconductors.

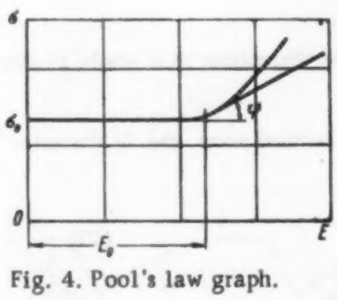

Fig. 4. Pool's law graph.

It is known from the physics of semiconductors that as long as the incremental velocity $\Delta \bar{v}$, acquired from an electric field $\bar{E}$ by an electron over the length of its free path, is small as compared with the velocity $\bar{v}$ of the electron's thermal movement, the duration t_f of its free path does not depend on the electric field strength.

The velocity increment during the time of the free path is

$$\Delta \bar{v} = \frac{e\bar{E}}{m} t_f. \tag{25}$$

Since t_f does not depend on E, the electron mobility

$$U = \frac{\Delta v}{E} = \frac{e}{m} t_f \tag{26}$$

does not depend on E either.

The current density is

$$\bar{j} = neU\bar{E} = \sigma \bar{E}, \tag{27}$$

where n is the current carriers' concentration; e is the electron charge; and σ is the semiconductor conductance.

Thus, the conductance does not depend on the field strength, and the current density is proportional to it as long as $\Delta v \ll v$ and the current carriers' concentration is constant, i. e., under those conditions the semiconductor follows Ohm's law. It is also naturally assumed that the semiconductor temperature does not vary with the electric field strength. Deviations from Ohm's law can occur at such low field strengths as 10^3-10^4 v/cm. The nature of these deviations depends both on the type of semiconductor and on the presence of impurities. At a field strength of the order of 10^4-10^5 v/cm the conductivity begins to rise according to Pool's law:

$$\sigma=\sigma_0 e^{\varphi(E-E_0)}. \tag{28}$$

The meaning of the parameters σ_0, φ and E_0 is made clear in Fig. 4. The voltage across the thermistor at the instant a pulse acts on it may reach a value of the order of several tens of volts. Since the distance between the conductors inserted into the bead is of the order of 0.1 mm or less, and the field between them is nonuniform, one should expect a field strength attaining values for which Pool's law is applicable.

Computations made on the basis of these assumptions lead to the following relation between the dc thermistor resistance due to the effect of a high field strength and the applied UHF power:

$$\Delta R_p(P)=\frac{2R_0\varphi k_1 U_0}{n}\left[\sqrt{\frac{P}{P_L}-1}-\cos^{-1}\sqrt{\frac{P_L}{P}}\right], \tag{29}$$

where R_0 is the thermistor resistance, and k_1, U_0 and P_L are constants.

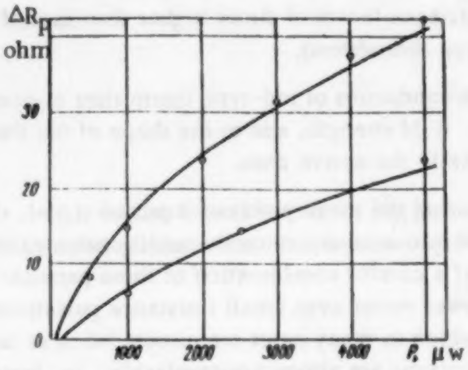

Fig. 5. Relation of ΔR_p to P calculated from (29). o-points obtained experimentally.

Figure 5 shows the graph of the relation between ΔR_p and P for two thermistors type T9, calculated from (29). (The constants were determined experimentally.) The same graph shows experimentally obtained points. It will be seen from Fig. 5 that the agreement between the theoretical and experimental data is good.

It should be noted that the resistance pulsations due to a high field strength are only characteristic for bead thermistors. In thermistors type TSh, whose lead-in conductors are spaced at 1 mm, the above effect was not observed.

Irreversible thermistor parameter changes (breakdowns). When a sufficiently high voltage is applied to a thermistor irreversible changes are observed in its parameters in addition to pulsations of its resistance. In the majority of cases a breakdown of the thermistor is accompanied by a rise in the power required to raise its resistance to a given value. Moreover, after a breakdown a drop in the amplitude of resistance pulsations is observed for the same pulse power and energy, as well as a variation in the UHF thermistor impedance for the same dc resistance. The above was ascertained by means of the following experiment.

A thermistor cell was matched to a UHF line for an unmodulated signal up to a voltage standing wave ratio ≤ 1.05. Next a pulsed UHF signal was fed to the line and its power was gradually raised until the thermistor breakdown (the breakdown was judged by a sharp deflection on the automatic power meter). Next, the voltage standing wave ratio was again measured for the same value of the dc thermistor resistance. After the breakdown the standing wave ratio rose to a value of the order of 1.5-2.

If the pulse power is raised above a certain limit for the same mean power the thermistor breaks down completely. In the majority of cases a sharp rise in the thermistor resistance up to several tens of kilohms is observed, but sometimes the lead-in conductors are shorted.

The value of the pulse power for which an irreversible change in the thermistor parameters occurs varies among individual thermistors and thermistor types. Thus, for bead thermistors it is in the range of 0.5-20 w, and for rod-type thermistors it is between 50 and 200 w.

Conclusions. The study of the thermistor characteristics under the effect of pulsed UHF signals has shown that under these conditions a thermistor behaves as a body with a small thermal time constant (of the order of tens or hundreds of microseconds), as the result of which the temperature of the thermistor (to be more precise, of its active zone) pulsates at the UHF pulse repetition frequency.

It has been established that the conductivity of a thermistor changes almost instantaneously under the effect of a relatively high voltage arising across it due to the action of a UHF pulse. The two phenomena lead to the thermistor resistance pulsating at the pulse repetition frequency of the signal.

The thermistors possess a limited electric stability and may change their parameters and even fail completely under the effect of UHF pulsed signals, although the mean power for the period of their action on the thermistors may be well below the overheating danger point.

The experimental study of various UHF thermistor types has shown that they can be divided by their reactions to a pulsed signal into two groups. Spherical bead thermistors with inserted lead-out conductors have a considerably lower electrical strength, and for the same pulsation conditions their resistance is tens of times higher than that of cylindrical thermistors with lead-in conductors to their butt ends (rod-type thermistors).

Such a difference is mainly due to the larger distance between the conductors of rod-type thermistors as compared with the bead type, thus providing a considerably smaller electric field strength, and to the shape of the thermistor and the greater spacing of its conductors, which increase substantially the active zone.

In designing and using thermistor power meters intended for measuring the mean power of a pulsed signal, the peculiarities of thermistor operation under such conditions must be taken into account, since the additional measurement errors may otherwise become excessively large. The importance of a careful consideration of these peculiarities is shown, for instance, by the fact that in the widely used VIM-1 power meter even small resistance pulsations, characteristic for rod-type TSh-2 thermistors, provide additional errors which in many cases are almost twice as large as the basic error of measurement. For such applications bead-type thermistors are altogether unsuitable. For instance, the use of such thermistors in substitution-type dc bridges for measuring the mean power of a pulsed signal may lead to errors of tens of percent [2]. It should be noted that the same bridges with thermistors of the TSh-2 or similar types have no additional errors as long as the pulse power does not attain values of the order of 100 w at which these thermistors break down.

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LOW-VALUE UHF FILM RESISTORS

G. M. Strizhkov

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For certain UHF measurements it is necessary to have low-value nonreactive resistors. It has been shown in [1] that film resistors in the shape of thin discs have a low reactance. The same work gives an expression for a transfer impedance of a coaxial line segment with a metallic diaphragm of thickness d placed perpendicularly to the line axis:

$$|\overline{Z}_{н}| = \frac{|\overline{U}_{н}|}{|\overline{I}_{н}|} \approx R_{н} \left| 1 - j \cdot \frac{d^{2}}{3\delta^{2}} - \frac{7}{90} \cdot \frac{d^{4}}{\delta^{4}} \right| \approx$$

$$\approx R_{н} \sqrt{1 - \frac{2}{45} \cdot \frac{d^{4}}{\delta^{4}}}, \tag{1}$$

where $\bar{U}_H$ and $\bar{I}_H$ are the voltage at the output and the current at the input of the film diaphragm; R_H is the dc resistance of the film, and δ is the depth of current penetration, which depends on frequency.

Equivalent electrical circuit for the resistors. On the basis of the current distribution along the film cross section (Fig. 1) it is possible to derive an equivalent circuit shown in Fig. 2. The skin effect is accounted for by inductance L_T, which increases with the thickness of the film. The shunting inductances may be considered as representing parallel connected elementary radial strips, with the total inductance being increased by deviations from symmetry [1]. By symmetry we understand the lack of an azimuth relation between the thickness and the resistance of the film.

Assuming that load current I flowing through the above resistor is $\ll I_2$, and that $R_1 = R_2 = 2R_H$ and $L_1 = L_2 = 2L_H$, we find

$$|\bar{U}_H| = |\bar{I}_H| \sqrt{\frac{R_H^2 + \omega^2 L_H^2}{R_H^2 + \omega^2\left(L_H + \frac{1}{4} L_T\right)^2}}. \tag{2}$$

It follows from (1) that for $|\bar{I}_H|$ = const the output voltage decreases with a rising frequency, since δ decreases. The same phenomenon is represented by (2) if we assume that $L_T \gg L_H$:

$$|\bar{U}_H| \approx |\bar{I}_H| R_H \frac{1}{\sqrt{1 + \omega^2\left(\frac{L_T}{4R_H}\right)^2}}. \tag{3}$$

Hence, thick films ($\delta_{min} \leq d$) and thin films ($\delta_{min} > d$) without asymmetry have a "falling" frequency characteristic, which is represented by the equivalent circuit for $L_T \gg L_H$. Such a characteristic obtained experimentally is given in Fig. 3.

It is difficult to make thin plates symmetrical. Deviations from symmetry increase considerably inductance L_H. For $L_H > L_T$ the output voltage rises with frequency:

$$|\bar{U}_H| \approx |\bar{I}_H| R_H \sqrt{1 + \omega^2\left(\frac{L_H}{R_H}\right)^2} \approx |\bar{I}_H| R_H (1 + af^2). \tag{4}$$

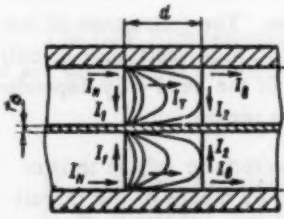

Fig. 1

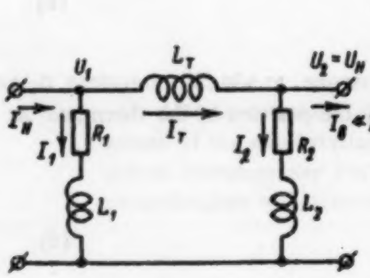

Fig. 2

Manufacture of low-value resistors. Surfaces can be covered with thin metallic films by various means. The simplest of these consists in burning silver onto a ceramic base.

As a base we used various ceramics and glass. Films covering glass do not adhere sufficiently well. Moreover, glass does not possess the required strength. Of the three types of ceramic, radio-porcelain, steatite and ultra-porcelain, it was found that steatite was the most suitable. Radio-porcelain retains securely the films, but it is very brittle and therefore cannot be used as a base. Ultra-porcelain is sufficiently strong and retains the films well, but it has flaws. The best results are obtained with compressed steatite. The surface of a selected base must first be polished until a smooth and even surface is obtained, and then washed and carefully degreased. The fired paste should be smeared in the proportion of 0.01 g per 1 cm² onto a revolving base (1400 rpm), since this tends to a uniform distribution of the layer of paste and in addition the layer is evened out by the centrifugal force. Moreover, any nonuniformity in the layer or the presence of impurities on the revolving surface of the baseproduces circles, which serve as indications of defects in the film.

The most important manufacturing stage consists in checking the film thickness and providing geometrical and electrical symmetry.

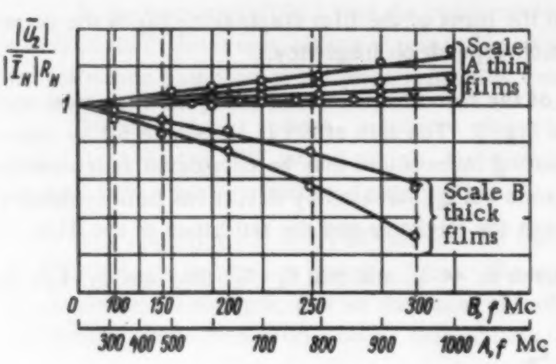

Fig. 3

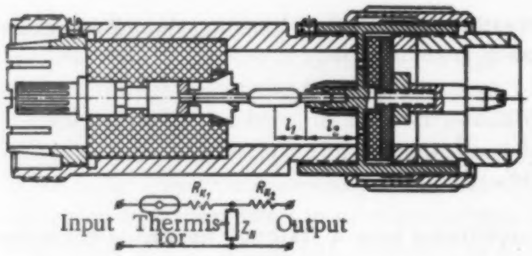

Fig. 4

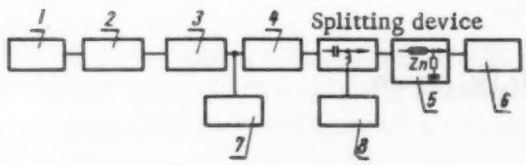

Fig. 5. Schematic of an instrument for determining the combined frequency characteristic of the resistor and thermistor. 1) Oscillator type GSS-12; 2 and 4) fixed 10 db attenuator type AS-1; 3) resonance filter; 5) thermistor type TSh-2B; 6) sensitivity-calibrated receiver; 7) tube voltmeter type VKS-7B; 8) reference thermistor bridge type MTO-1.

The film thickness $\underline{d}$ can be determined by measuring the mass of the base before and after firing, with a subsequent computation from the mass and area of the deposited silver, or by the dc resistance of the film.

The latter method assumes a uniform distribution of the silver film. Any discontinuities in the film raise its resistance, leading to errors in estimating its thickness. For thicknesses of the order of 1μ both methods showed satisfactory agreement.

The uniformity of the deposited film thickness and of its electrical resistance was checked by means of a probe with four needle contacts placed along the radii of the film. The two outside needles served to carry the current and the two inside ones to measure the potential. The needles were connected to the appropriate contacts of a dc bridge for measuring small resistances. By moving the probe and measuring corresponding resistances it is possible to obtain a topographic representation of the film thickness. Films which had deviations from the mean thickness not exceeding ± 5% were considered satisfactory. In addition, direct measurements of the differences of radii $R_0 - r_0$ (Fig. 1) were also made.

Properties of low-value resistors. Experience has shown that adherence to the above measures improves the frequency characteristics of resistors; however, the soldering of contacts to the film may produce defects which it is very difficult to discover. Therefore, the frequency characteristic of each completed resistor was determined by means of a thermistor bridge and a receiver. In this circuit the current-measuring thermistor must be placed in the immediate vicinity of the measured resistor. Figure 4 shows the construction of the thermistor head intended for measuring low-value film resistors. The resistor has concentric contacts soldered directly to the deposited silver layer and fitted with threaded connections to the incoming and outgoing lines. The resistances of the threaded connections were placed in high-resistance circuits, thus making the output resistance of the head only dependent on the voltage drop across the film resistor.

The overall frequency characteristic of the resistor and the thermistor was determined by means of the circuit shown in Fig. 5:

$$|\overline{U}_{\kappa}| = R_{\kappa}\sqrt{\frac{P_T}{R_{T_0}}}\,(1+D_0 f^2) = |\overline{U}_0|\,(1+D_0 f^2), \tag{5}$$

where P_T is the power measured by the thermistor; R_{T_0} is the thermistor dc resistance, to which corrections Δ_1 and Δ_2 were applied, which accounts for the thermistor frequency characteristic and the position of the thermistor at a distance l_{eq} from the resistor.

It was shown in [2] that

$$\Delta_1 = -\frac{1}{2}\omega^2 R_{T_0}^2 C_s^2, \tag{6}$$

where C_s is the equivalent capacitance shunting the thermistor bead, equal to 0.3 μμf for a thermistor type TSh-2B [2].

The values of l_{eq} and Δ_2 are calculated from the construction of the thermistor head, which consists of two sections with characteristic impedances Z_{01} and Z_{02} and lengths l_1 and l_2. In the thermistor head used by us l_{eq} = 12 mm. Thus, parameter a of the frequency characteristic (4) is equal to

$$a = D_0 - 2\pi^2 R_{T_0} C_s^2 - 2\pi^2 \frac{l_{eq}^2}{C^2}, \tag{7}$$

where $C = 3 \cdot 10^{11}$ mm/sec.

Coefficient D_0 was determined by measuring voltage $|\bar{U}_H|$ at various frequencies and a constant power P_T by means of a receiver. In order to improve the waveform of current I_H a resonance filter was included in the circuit and tuned by means of a tube voltmeter.

These measurements have shown that the output voltage $|\bar{U}_H|$ has for thin films and a constant $|\bar{I}_H|$ a square law relation to frequency, thus confirming the assumed equivalent circuit and computations. For low-value resistors in the range of $R_H = (0.5\text{-}3) \cdot 10^{-3}$ ohm coefficient $a = (0.05\text{-}0.25) \cdot 10^{-6}$.

The relation between the resistance of the low-value resistors and the current flowing through them was also determined. It has the form of

$$\Delta R_H = (b I_H^2)\%. \tag{8}$$

Coefficient b depends on the value of R_H. For an $R_H = (1\text{-}5) \cdot 10^{-3}$ ohm, $b \approx 7 \cdot 10^{-4}$.

It will be seen from (8) that in measuring the dc resistance R_H it is possible to use current of the value of $I_H \leq 3$ amp. This requirement is met by bridge MOD-54 and a potentiometer circuit.

In order to determine the stability with time of these film resistors, five were measured each day over a period of six months. The resistance varied in the first month by about 1.5%, for the second by 0.5%, and for subsequent months by 0.2%. The greatest deviation of the resistance is observed in the first 5 days.

Measurements of the temperature coefficient of film resistors have shown that its mean value amounts to 0.28% per degree C.

Conclusion. The application of the technique described above for determining coefficient a of the frequency characteristic and the use of appropriate corrections for the effect of the external temperature and for the aging of resistors made it possible to determine the value of $|\bar{Z}_H|$ with an error not exceeding $\pm(2\text{-}3)\%$ at a frequency of 1000 Mc.

The UHF film resistors made in the manner described above have values between $0.1 \cdot 10^{-3}$ and $10 \cdot 10^{-3}$ ohm, and are suitable for use in instruments measuring small voltages at radio and ultrahigh frequencies.

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All abbreviations of periodicals in the above bibliography are letter-by-letter transliterations of the abbreviations as given in the original Russian journal. *Some or all of this periodical literature may well be available in English translation.* A complete list of the cover-to-cover English translations appears at the back of this issue.

METHOD FOR CHECKING SCALERS

V. P. Rynkevich

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 54-55, March, 1961

To date there is no reliable method for checking scalers used in conjunction with radiation detectors in nuclear physics for counting the number of particles. The accuracy of their counting is checked by feeding the instrument with regular pulses which have a known repetition frequency over a set period. The number of pulses the instrument is supposed to count is known with great accuracy, and the instrument's resolution time is determined by the spacing between two adjacent counted pulses. However, the conditions of counting in this case do not correspond at all to the working conditions of the instrument, since in fact the instrument does not count pulses with a regular, but with a random, distribution with respect to time. It is obvious that checking the instrument by means of regular pulses cannot be considered satisfactory, since with a random variation of the intervals between pulses the probability of the instrument missing pulses varies and differs from the probability in the case of regular pulses. In order to meet this condition scalers are now being checked by means of generators which produce trains of double and treble pulses, thus making it possible to evaluate more accurately the scaler's dead time. However, in this case the test conditions differ considerably from the scaler's working conditions.

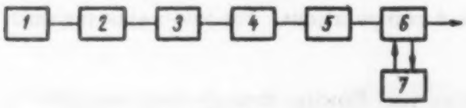

Fig. 1.

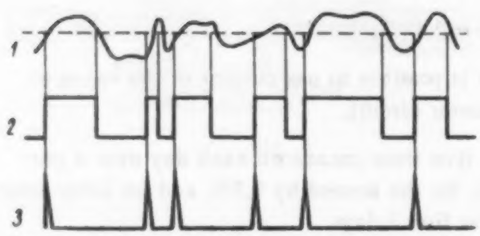

Fig. 2. 1) Noise voltage at the output of the video amplifier (the dotted line shows the level of the Schmitt trigger operation); 2) voltage at the Schmitt trigger output; 3) voltage at the generator output.

The most accurate checking method would consist in feeding to the scaler input pulses distributed in time according to the same law as the pulses of a radiation detector (for instance, a scintillator or a Geiger-Mueller counter). It is known that in this instance the following distribution law holds:

$$P(T)=\frac{1}{\overline{T}}e^{-\frac{T}{\overline{T}}}, \qquad (1)$$

where T is the pulse repetition period, and $\overline{T}$ is the mean repetition period.

The circuit described below produces the required pulse sequence [1]. The block schematic of the device is shown in Fig. 1, and the shapes of voltages across its components in Fig. 2.

The noise source 1 (Fig. 1), consisting of an ordinary noise diode, produces a noise voltage which has a wide frequency spectrum and is fed to video amplifier 2 with an adjustable bandwidth. The output voltage of the video amplifier has the form of a random process with a spectrum determined by the amplifier bandwidth. This voltage operates the Schmitt trigger 3. The trigger operates when the noise voltage crosses its tripping level. Following differentiation (device 4) and limitation of trigger pulses, a sequence of short, single-polarity pulses with a statistical distribution with respect to time is obtained.

The distribution of the output pulses according to the required law is determined by the following circumstances.

The noise voltage produced by the diode represents a random stationary process, which follows a normal distribution law, i. e., the distribution function of the noise voltage amplitude, and that of derivatives of any order follow a normal distribution law. With a correctly chosen operating condition the video amplifier constitutes a linear system for the noise voltage fed to its input. Since any stationary random process is transformed by a stationary linear system into a stationary process, the voltage at the output of the video amplifier will also represent a stationary process with another amplitude distribution law, and the spectrum of this process will be limited by the bandwidth of the amplifier.

The probability of obtaining a pulse at the output of the Schmitt trigger, whose input is fed by a noise voltage from the video amplifier, is equal to the probability of a random stationary process crossing a given level. The latter

probability is a constant quantity and may be calculated from the formula

$$P(x_0) = W(x_0) \int_0^{\infty} W(y)\, dy = \text{const}, \tag{2}$$

where W(x) is the distribution function of a random stationary process; W(y) is the distribution function of the derivative of a random stationary process; x_0 is the crossing level.

Formula (2) covers crossings in the upward direction. For crossings in the downward direction the integral is taken for negative values of the derivative.

The mean value of crossings per unit time is also a constant quantity, determined from formula

$$\nu = \frac{1}{T} \int_0^{\infty} \int_t^{t+T} y W_2 (x_0, y, t)\, dy\, dt. \tag{3}$$

For a sufficiently large time interval T the number of level crossings in the upward direction will be

$$n = \nu T. \tag{4}$$

These events occur in a random manner and independently from each other. The probability that k out of n events will occur in a set time interval τ is calculated from the formula

$$P_n(k) = \frac{\tau}{T} = \frac{\nu\tau}{n}. \tag{5}$$

In this case, in view of a limited spectrum of noise voltages and finite operating time of the circuit, the mean number of pulses appearing in a unit of time will be small in relation to the number of tests and the total number of pulses. In other words, the characteristic of (5) consists of a large number of tests n and a small probability of an event occurringduring one test. Under such conditions it is possible to apply Poisson's formula:

$$P_n(k) = \frac{(\nu\tau)^k}{k!} \cdot e^{-\nu\tau}. \tag{6}$$

It is known from the theory of random processes that the intervals between events in this case are distributed according to law (1). Hence, by differentiating the pulses produced by the Schmitt trigger and passing them through a limiter 5 (Fig. 1) so that at its output only positive pulses appear, it is possible to obtain the required distribution of the output pulses with respect to time. The result thus obtained is independent of amplitude distribution of the process (Gaussian, Rayleigh or any other distribution), providing the process is stationary.

It will be seen from (2) and (3) that the mean pulse repetition frequency is determined by the noise voltage spectrum at the input of the Schmitt trigger and the operating level of the circuit. Thus, by changing the amplifier bandwidth or the operating level of the Schmitt trigger it is possible to vary the mean pulse repetition frequency.

In order to obtain a predetermined number of pulses the generator is equipped with a reference scaler circuit 7 (Fig. 1) with a capacity equal to the required number of pulses. As soon as the full capacity of the scaler circuit is reached the generator is automatically cut out. The circuit is coupled to a commutating stage 6 (Fig. 1).

By means of the above device it is possible to check scalers with an accuracy determined by that of the reference counter.

When an ordinary Schmitt trigger circuit is used it is possible to check scalers with a time resolution of 1 μ sec or more.

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REFERENCE TRIMMER DISC CAPACITOR FOR FREQUENCIES UP TO 200 Mc

A. L. Grokhol'skii

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 55-57, March, 1961

The high-frequency disc capacitor KVCh-8 developed by the NGIMIP (Novosibirsk State Institute of Measures and Measuring Instruments) has been used successfully in several USSR scientific research institutes. The capacitor has been recently improved and thoroughly investigated in order to extend its use for testing and research purposes. The main advantage of this capacitor consists in the small extent to which its effective capacitance depends on the operating frequency. At frequencies up to 20-30 Mc and measurement errors of the order of 0.5-1% its capacitance can be considered independent of frequency and equal to its capacity at audio frequencies. For higher frequencies or when more accurate measurements are required, it is necessary to apply to the low-frequency capacitor readings certain corrections which can be easily calculated from the formulas given below.

Electrical circuit of the capacitor. The capacitor consists of two disc plates at a distance d from each other in such a manner that their axes of symmetry coincide. The capacitor conductors are connected to the midpoints of its plates.

The disc capacitor can be represented by an equivalent circuit consisting of a constant capacitance C_0 connected in series with the residual inductance L_ω of the capacitor and its shunt conductance g (Fig. 1).

The effective capacitance can be determined from the formula

$$C_e = \frac{C_0}{1-\omega^2 L_\omega C_0}. \tag{1}$$

The value of the residual inductance L_ω of the capacitor is calculated from its geometrical dimensions. The formula for determining L_ω was first derived by R. King and then reproduced by N. A. Kouzov. They proved that the residual inductance is determined by the distance between the plates and is equal to

$$L_\omega = \frac{d}{2} 10^{-9}\,\text{h}. \tag{2}$$

This formula (d, cm) holds for frequencies at which radius a of the capacitor discs is considerably smaller than the wavelength.

In the case when the wavelength and the dimensions of the capacitor plates are of the same order, formula (2) no longer holds. In such a case it is better to use the formula derived at the NGIMIP:

$$L_\omega = \frac{d}{2}\left[1+\frac{1}{3}\pi^2\left(\frac{a}{\lambda}\right)^2\right]10^{-9}\,\text{h}. \tag{3}$$

Formulas (1) and (3) do not account for inductance L_0 of the capacitor connections, without which it is impossible to use it in a measuring circuit. If this inductance is considered to be connected in series with the residual inductance L_ω the effective capacitance can be calculated from the formula

$$C_e = \frac{C_0}{1-\omega^2 C_0 (L_\omega + L_0)}. \tag{4}$$

Normally the total inductance $L_\omega + L_0$ is mainly determined by the value of L_0, since L_ω is always much smaller than L_0. For this reason special attention should be paid, in designing disc capacitors, to reducing the value of L_0. This can only be accomplished by making the capacitor connections with a minimum inductance. In order to fulfil this requirement several designs of disc capacitors have been produced. The most successful design was that of capacitor KVCh-8M, the relation of whose capacitor temperature coefficient to the gap between the plates, i. e., to the set capacitance, is shown in Fig. 2.

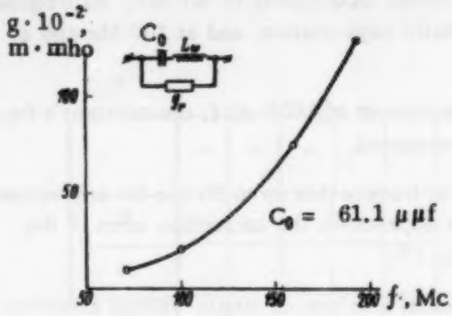


Fig. 1.

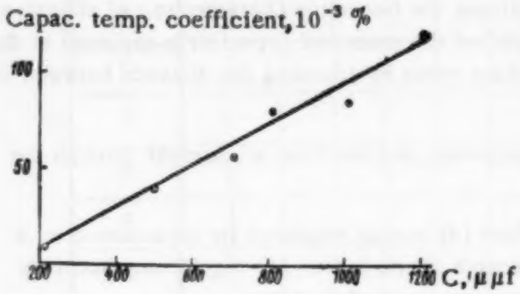


Fig. 2.

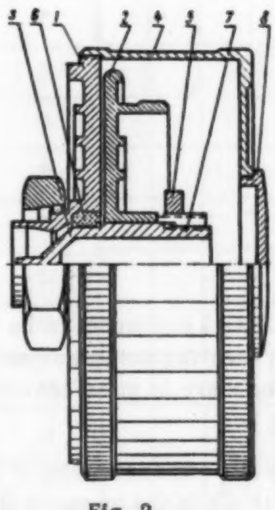


Fig. 3.

25-1250 $\mu\mu$f trimmer disc capacitor KVCh-8M is shown in Fig. 3. It consists of a massive brass base 1, which has several circular ribs to provide it with the required rigidity and which serves as the low-potential plate of the conductor. On one side the base is fitted to a connection which has the dimensions of a standard 75-ohm coaxial socket, and on the other it is fitted by means of a ring steatite insulator 6 to the central axial contact 3.

The upper high-potential plate 2 also has stiffening ribs. It is tightly fitted over the axial contact, but it can be displaced along the contact in its rotational movement. The distance between the plates is varied by means of screw pair 7. The position of the plates is secured and the contact resistance decreased by means of nut 5. The high-potential plate is screened by means of casing 4, whose bottom has a hole which serves for accurate adjustments and is closed by stopper 8.

In order to eliminate any possible deformation in the operation of the capacitor, all its parts were subjected before assembly to a 3-hour heat treatment at a temperature of 300-400°C with subsequent cooling in the oven. After annealing the working surfaces of plates 1 and 2 are lapped on a cast iron plate. It is advisable to cover all the capacitor components with a bright layer of chromium. These surfaces should then be cleaned only by chemical means. The capacitor assembly is started by first soldering together components 1, 3 and 6 with a low melting point solder (120-150°C). All the components are assembled finally in a hot state, when the solder becomes fused, and the working surfaces 1 and 2 are pressed tightly together and secured by nut 5. In this condition the capacitor is allowed to cool gradually.

Only when all these requirements are fulfilled is it possible to produce a good capacitor with a range of 25 to 1250 $\mu\mu$f and plate diameters of 100 mm.

The minimum capacitance of the trimmer with plate 2 and screen 4 removed amounts to 8 $\mu\mu$f. This capacitance is mainly concentrated in the area of the ceramic ring insulator 3. The edge capacitance between the moving plate 2 and casing 4 over the whole trimmer range (25-1250 $\mu\mu$f) remains practically constant at 6.3 $\mu\mu$f. It is due to the field between the edge of the moving plate and the screen. The capacity between the screen and the plane of the moving plate changes very little, remaining in the range of 1.5-2.5 $\mu\mu$f.

All the above-mentioned residual capacitances are considerably smaller than the main capacitance.

The relation shown in Fig. 2 holds for a capacitor whose components 1, 2, and 3 (see Fig. 3) are made of brass. It is possible, if necessary, to reduce the capacitance temperature coefficient by making the above components in different metals with appropriate coefficients of linear expansion, which would provide the required temperature compensation.

The losses in the capacitor were determined in the frequency range of 70 to 200 Mc. These losses and their variations in terms of a conductance shunting the capacitor are shown in Fig. 1.

The inductance of the capacitor's connection was obtained by working out the results of a number of measurements of the effective trimmer capacitance at several frequencies, including 200 Mc, by means of an accurate twin-T admittance bridge. The value thus obtained amounted to $L_0 = 1.92 \cdot 10^{-9}$h $\pm$ 1%.

Thus, it is possible to use capacitor type KVCh-8M in the range from audio frequencies to 200 Mc. Its frequency correction at 20 Mc and a capacity of 1200 $\mu\mu$f amounts to 2.2% of its static capacitance, and at 100 Mc and a capacitance of 100 $\mu\mu$f it amounts to only 5.5%.

It should be noted that at frequencies of the order of 200 Mc and capacitances of 1000 $\mu\mu$f, the trimmer's frequency error was not determined, since in practice such values are not encountered.

The error in setting the effective capacitance of trimmer KVCh-8M at frequencies up to 30 Mc for any values of static capacitance amounts to less than 0.2-0.5%. At higher frequencies it depends on the correction term in the denominator of (4); if this term is smaller than 20% the error will not exceed 1%.

Conclusions. The above disc capacitor can be used for various purposes as a reference single-valued effective capacitance, or as a many-valued capacitance set at will to any value within its range, for testing or research purposes.

The KVCh-8M capacitors are particularly convenient for determining the frequency characteristic of effective capacitances of other capacitors by the substitution method. In this method the measured capacitor is replaced at the high frequency by the KVCh-8M capacitor, which is set to the appropriate value by adjusting the distance between its plates.

The preliminary adjustment is made with the capacitor screen removed, and the final adjustment through the opening in the screen.

The actual capacitance of the disc trimmer is then calculated from (4) having measured its capacitance on a low-frequency bridge. Such a method of checking capacitors is very simple, since it does not require any accurate high-frequency capacity measuring devices. It is quite sufficient to have any type of Q-meter.

USE OF AUXILIARY TABLES IN CHECKING INSTRUMENT 26-I

A. F. Kamnev

Translated from Izmeritel'naya Tekhnika, No. 3,
pp. 57-59, March, 1961

Checking the duration and delay of pulses in instrument 26-I by means of the UPIG-2 equipment is the most labor-consuming operation of all those required for testing the above instrument. In checking the instrument for these two parameters it is necessary to make over 60 measurements of frequency f_r and make as many conversions into time intervals τ.

The checking of instruments by means of time parameters can be greatly simplified and accelerated if use is made of a table of frequency tolerances which correspond to a certain number of markings in the measured time interval.

We have compiled such tables for instrument 26-I, whose pulse duration varies from 0.2 to 10 μsec with a tolerance of $\pm$(10% + 0.025 μsec) and delay time varies from 1 to 21 μsec with a tolerance of $\pm$(10% + 0.5 μsec).

The number of markings which can be placed in the measured time interval is chosen according to the operation instruction No. 151. At the intersection of the horizontal line which passes through the index which corresponds to the value of the measured time interval, and the vertical line which passes through the index corresponding to the number of markings in the measured time interval, two numbers are placed, one above the other. These numbers indicate the reference oscillator frequencies in kc corresponding to the tolerances for the measured time interval.

It is convenient to present the above tables in the form of a two-sided board with a cursor for ease in taking readings.

TABLE 1. Auxiliary Table for Checking Pulse Delay Time in Instrument 26-I

n τ	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25	26	27	28	
1	5000 1250	7500 1875																										1
2	1540 740	2310 1110	3070 1480	3850 1850																								2
3		1360 790	1820 1050	2280 1315	2720 1580	3180 1840																						3
4			1290 815	1610 1020	1935 1225	2260 1430	2580 1632	2900 1835																				4
5				1250 830	1500 1000	1750 1160	2000 1334	2250 1500	2500 1670	2760 1832																		5
6					1223 845	1430 987	1633 1127	1837 1266	2040 1410	2250 1550	2450 1690	2660 1832																6
7					1035 735	1210 854	1380 976	1550 1097	1725 1220	1900 1342	2070 1460	2240 1590	2415 1710	2590 1830	2760 1950													7
8						1045 750	1190 860	1345 970	1494 1075	1642 1180	1792 1290	1940 1400	2090 1506	2240 1612	2390 1720	2540 1850	2690 1955											8
9							1050 770	1180 865	1315 962	1450 1057	1580 1150	1710 1250	1845 1345	1975 1442	2100 1535	2240 1635	2370 1730	2500 1825	2640 1925									9
10	235 174							1060 780	1175 870	1295 955	1415 1040	1530 1130	1650 1220	1768 1305	1884 1390	2000 1480	2120 1564	2250 1652	2350 1740	2480 1830	2590 1910							10
11	213 158.5								1064 807	1170 875	1280 954	1380 1030	1490 1110	1595 1190	1700 1278	1810 1350	1915 1430	2020 1508	2125 1588	2240 1670	2340 1745	2450 1825						11
12	192.5 146									1067 804	1165 879	1260 950	1360 1020	1455 1095	1555 1168	1650 1240	1750 1315	1845 1385	1945 1460	2040 1532	2130 1605	2230 1678	2330 1755					12
13	178.5 136										1070 810	1160 880	1250 947	1340 1015	1430 1080	1515 1149	1605 1215	1696 1285	1785 1315	1875 1416	1965 1435	2050 1555	2142 1640	2230 1690	2320 1755			13
14	165 125.5	248 189										1075 817	1156 882	1240 943	1323 1005	1405 1070	1488 1131	1570 1195	1655 1257	1735 1320	1818 1383	1900 1445	1984 1510	2062 1570	2150 1635	2230 1695		14
15	154 118	231 176.5											1075 824	1155 884	1230 943	1310 1000	1384 1060	1460 1120	1540 1178	1615 1235	1691 1292	1770 1352	1845 1412	1925 1470	2000 1530	2075 1587	2150 1643	15
16	144 110.6	216 166												1078 830	1150 884	1223 940	1295 995	1366 1050	1438 1105	1512 1160	1582 1215	1655 1270	1725 1328	1800 1382	1873 1438	1945 1490	2010 1547	16
17	134 104	201 156													1080 834	1150 885	1215 940	1283 990	1350 1040	1480 1093	1488 1145	1555 1200	1620 1250	1688 1303	1755 1355	1825 1406	1890 1457	17
18	127.5 98.5	191 148	255 197													1082 840	1147 886	1210 935	1274 985	1340 1034	1400 1085	1465 1132	1530 1182	1592 1232	1655 1283	1720 1330	1784 1380	18
19	120.4 93.5	181 140	241 187														1085 842	1143 887	1205 935	1264 980	1325 1028	1385 1074	1445 1123	1505 1168	1566 1216	1628 1262	1686 1310	19
20		171 133.6	228 178															1085 845	1143 890	1200 934	1255 980	1313 1020	1372 1066	1428 1110	1485 1155	1543 1200	1600 1243	20
21		163 127	216 169.5																1086 850	1140 890	1196 933	1250 975	1305 1016	1360 1080	1414 1102	1468 1144	1520 1188	21
	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25	26	27	28	

TABLE 2. Auxiliary Table for Checking Pulse Durations in Instrument 26-I

τ \ n	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	
0.1	30800 14800																			0.1
0.2	12900 8160	19350 12230	25800 16340																	0.2
0.3		12230 8450	16340 11280	20400 14070	24500 16900	28600 19700														0.3
0.4			11940 8600	14900 10770	17900 12900	20830 15050	23800 17200	26800 19350												0.4
0.5				11770 8700	14100 10430	16450 12170	18800 13920	21200 15650	23500 17400	25900 19150										0.5
0.75						10780 8230	12300 9420	13840 10600	15400 11760	16950 12950	18200 14300	19900 15400	21400 16600	22900 17750	24400 18950					0.75
1.0	2290 1780							10300 8050	11300 8900	12430 9770	13650 10650	14850 11550	16000 12430	17150 13350	18300 14200	19450 15100	20600 16000	21700 16900	22900 17750	1.0
1.5	1508 1194	2265 1790										9820 7750	10550 8350	11300 8970	12000 9550	12850 10150	13600 10750	14350 11320	15100 11950	1.5
2 0	1128 900	1690 1348	2260 1800															10700 8550	11270 9000	2.0
2.5		1346 1083	1795 1455	2240 1805																2.5
3.0		1170 902	1490 1205	1865 1505	2240 1805															3.0
3.5			1258 1050	1570 1310	1887 1570	2130 1835														3.5
4.0			1088 910	1360 1135	1630 1335	1900 1585	2170 1810													4.0
4.5				1225 1008	1470 1210	1710 1410	1955 1600	2200 1815												4.5
5.0				1110 910	1340 1090	1555 1270	1780 1455	2000 1635	2220 1820											5.0
6.0					1110 910	1295 1060	1480 1210	1670 1365	1850 1516	2040 1670	2220 1820	2410 1970								6.0
7.0						1110 910	1270 1040	1430 1130	1590 1300	1750 1430	1910 1560	2030 1690	2220 1820	2320 1950						7.0
8.0							1110 910	1250 1020	1390 1135	1530 1250	1670 1365	1805 1480	1950 1590	2080 1750	2220 1830	2360 1955				8.0
9.0								1110 910	1235 1010	1360 1110	1480 1210	1605 1310	1730 1415	1855 1515	1930 1616	2100 1716	2220 1820	2350 1920		9.0
10.0									1110 910	1220 1000	1335 1090	1445 1180	1555 1270	1650 1365	1780 1450	1890 1545	2000 1640	2110 1725	2220 1820	10.0
	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	

MATERIAL RECEIVED BY THE EDITORIAL BOARD

CHECKING THE QUALITY OF INSTRUMENT PRODUCTION

P. M. Larionov

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 60, March, 1961

One of the most important activities of the Chelyabinsk GKL (State Inspection Laboratory) consists in checking the quality of production in instrument-making plants of the Chelyabinsk region.

In 1960 the GKL exerted great efforts to improve the checking of instrument production. In the first place responsibilities were allocated. The laboratory administration assigned to this work the most qualified personnel who had already been engaged for a considerable time in checking the quality of gauges and instruments.

Taking into consideration the necessity for a deep study of the causes of production defects, of the technology of production, the quality of raw materials, etc., all the inspection workers studied at the plants the entire technological production cycle right from the reception of the raw materials to the final assembly of instruments.

The laboratory established relations with all the scientific research institutes and design bureaus of the Chelyabinsk region which were engaged in the development of new instruments, and acquainted them with the procedure of state testing of new gauges and instruments, and established contact between them and the scientific research institutes of the Committee which carry out these tests.

It should be noted that the GKL has encountered difficulties in carrying out instruction 2-59 on coordinating the technical proposals for developing instruments with the organizations responsible for their design. There is, as yet, no definite criterion for determining which instruments should be subjected to state testing in the Committee's agencies. Therefore, it is often difficult for the GKL to determine which of the instruments should have their specifications developed in cooperation with the design bureaus.

On the basis of systematic testing of mass-produced instruments the GKL made a critical evaluation of the gauges and instruments in production, and made proposals for modernizing some of the instruments. On the basis of these suggestions the Sovnarkhoz (Council of National Economy) laid down certain dates by which the instrument-making plants have to submit modernized types of potentiometers, bridges, and slide-gauges for state testing, in order to start their production in 1961.

The GKL has considerably extended the volume of its work in testing instruments. Previously its activity was restricted to testing mass-produced articles, but in 1960 the laboratory undertook on behalf of the Committee the state testing of newly designed instrument models.

However, serious difficulties are being experienced in this work, due to insufficient laboratory personnel and lack of systematic information on the latest measurement techniques.

An important part of the work in checking instrument production consists in a constant study of the operational properties of gauges and measuring instruments produced in the plants of the region.

The laboratory has organized supervision of instruments at plants where a large number of instruments are used. Thus, the automatic electronic potentiometers and bridges of the "Teplopribor" Plant are being studied at a factory which uses a large number of these instruments.

The study of the operational properties of stop-watches made by the Zlatoust Horological Plant was organized in a different way. A consignment of stop-watches was taken from the stores of this plant and sent to four enterprises which studied them according to a program devised by the GKL. It was also specified that in case of a failure of the stop-watches they should be sent to a certain repair shop, which had received instructions on recording the type of repairs made, listing the repaired parts and the nature of the defect.

In addition to studying the properties of instruments at the plants of the region, the laboratory also uses the material received from other GKLs which studied the behavior of instruments made in the Chelyabinsk region.

After having been analyzed this material is brought to the notice of the directors of instrument-making plants. At the same time the plants received suggestions how to eliminate the defects discovered in the course of these studies.

Discussion at technical conferences with the workers of instrument-making plants of the technical data obtained in checking instruments, a careful planning of measures for improving production and quality control of instruments, produced good results.

In certain instances the GKL has also taken other measures. In particular, one of the plants was forbidden to deliver its slide-gauges to customers, and the OTK (Technical Control Division) of the plant was instructed to re-test all the slide-gauges in the stores and to carry out the final testing only after they had been stored there for two days.

The plant was also warned that if the defects in their slide-gauges were not speedily eliminated and the inspection made by their OTK improved, sterner measures would be adopted.

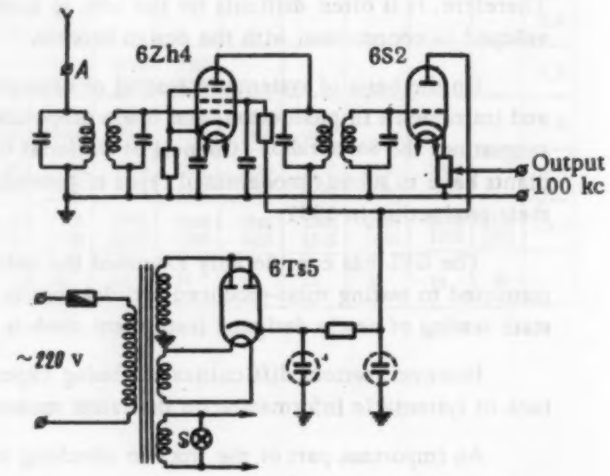
The Chelyabinsk GKL is continuing to improve its methods in checking instrument production. It is necessary to eliminate certain important failings in this work, such as an insufficiently profound study by the GKL personnel of the production technology of instruments, and of the reasons for defects in the manufactured gauges and instruments.

USING THE FREQUENCY OF 100 kc TRANSMITTED BY RADIO

N. F. Semenyuta

Translated from Izmeritel'naya Tekhnika, No. 3, p. 61, March, 1961

The use of the reference frequency of 100 kc at test points is impeded owing to the absence of special receivers. We propose a simple receiver for obtaining the frequency of 100 kc.



The receiver circuit (see figure) uses a 6Zh4 tube as a high-frequency amplifier and a 6S2 tube as a cathode follower circuit.

The grid and anode tuned circuits consist of intermediate-frequency transformers of an AR3 radio receiver, which are easily retunable to a frequency of 100 kc. The power transformer is also taken from the same receiver.

The signal obtained at the output of the cathode follower can be used instead of the 100 kc obtained from the crystal oscillators in the existing equipment, or else for checking the frequency of other oscillators.

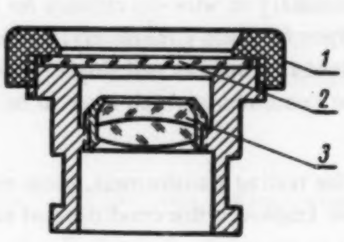
PROTECTIVE GLASS FOR AN EYEPIECE

A. P. Nikonorov

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 61, March, 1961

The outside surface of an eyepiece, which is not protected during operation, is the part of optical mechanical instruments which gets dirtiest under workshop conditions. The clarity of vision is disturbed, causing eyestrain and reducing productivity. This compels the operator, without the aid of a mechanic, to clean the lens himself or to blow the dust off its surface. Such "cleaning" at first damages the protective layer and then the lens itself. The damaged lens surface begins to corrode, becomes opaque and diffuses light. The instrument is thus put out of action for a considerable time, since optical repairs are complicated and require special equipment and qualified personnel. Often, owing to the lack of optical spare parts the instrument is written off as worn out and unserviceable.

The author suggests that a piece of photographic plate glass 1 mm thick be used as a protective covering and placed under the eyepiece holder.



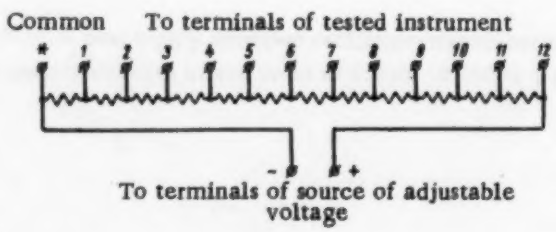
Protective glass 2, cut to the size of the internal diameter of the eyepiece holder 1, and carefully washed and cleaned (at the same time as the optical parts of the instrument), is placed under the eyepiece holder, which is then screwed onto the eyepiece (see figure). In order to improve the optical properties of the protective glass it is recommended that it be boiled in a 2% solution of NaOH for 1 hour, allowed to cool in the solution and then boiled in a 0.5% solution of acetic acid for 20-30 minutes, and then allowed to gradually cool in the solution. The quality of this layer will depend on the grade of glass and the time taken for its treatment. However, the brightening of the glass is not indispensable; it is sufficient to wash and clean it properly. The protective glass prevents the eyepiece lens 3 from being touched by hand and protects it from dirt and mechanical damage. The simple fixing of the protective glass makes it possible to change it without stopping work. In certain instruments the eyepiece holders have insufficient threads. For such instruments new holders should be made, which is not difficult. In the newly produced instruments the designers should provide eyepiece holders suitable for fixing protective glasses in them.

CHECKING THE QUALITY OF RECORDINGS IN THERMAL INSTRUMENTS

A. D. Snagovskii

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 62, March, 1961

In repairing and checking electronic potentiometers type ÉPP-09 as well as other thermal multi-channel recording instruments, it is necessary to check the operation of the switch and the recording device and also the quality of recordings. For this purpose it is recommended to use, in addition to the source of controlled voltage (IRN-50, IRN-53, etc.), yet another potential divider (see figure).



The attachment is made of 12 resistors of 4 ohm each, connected in series. The resistors are wound with 0.3 mm PÉSh-OMT wire on ebonite formers of 5 ohm resistors, and are placed in a 50 × 80 × 150 mm cabinet, whose front panel carries 13 spring clips. The attachment can be used for checking 3-6- and 12-channel instruments with different calibrations. For checking bridges a similar attachment can be made.

A voltage corresponding to the top scale reading is set on the divider and connected to the attachment, and the terminals of the various channels of the instrument are connected respectively to terminals 1-2 . . .12 of the attachment (the negative sides of the channels are connected to terminal 0); thus a recording is obtained for all the channels.

IMPROVEMENTS IN TESTING EQUIPMENT

D. I. Mogilevskii

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 62, March, 1961

The workers of the Irkutsk GKL (State Inspection Laboratory) are developing and bringing into use various new technical methods and devices which provide higher productivity of labor and reduce the time used in testing.

Technical inspector Leonchenko has made a switch for changing supplies in the UPV-5 set. This switch simplifies the testing of pointer instruments of different ranges.

Engineer Yakovlev has designed a switch for set UPS-1 which makes it unnecessary to wire-up circuits for each type of testing. A special attachment developed by him makes testing of heterodyne frequency meter crystal oscillators possible. He has also developed and is producing a portable set for testing electrical instruments and single-phase electricity meters in the place of their operation, thus reducing to a minimum the instruments which have to be sent to the laboratory for testing.

Moreover, the laboratory personnel have produced a high-tension chamber for testing transformers, have made an instrument, and mastered the technique, of checking grade II standard cells, have improved the conditions of storage batteries, etc.

The Irkutsk GKL renders technical assistance to factory test laboratories in improving their technique for checking production processes, and in applying the most up-to-date methods for inspecting measuring equipment.

Several factory laboratories have produced, according to recommended drawings, devices for checking large-size end gauges, the parallelism of planes in micrometers exceeding 100 mm, and for estimating the measuring efforts of micrometers. They have also produced, according to a circuit developed by the Irkutsk GKL, portable installations for checking electrical instruments.

The laboratory is planning other measures for developing technical progress in the sphere of measurements.

FROM FOREIGN JOURNALS

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 63, March, 1961

FEINWERKTECHNIK

No. 7, 1960

P. Heile and W. Deutsch. Modern techniques for measuring surface roughness. Existing methods and systems E and I are described as well as the high-precision instrument "microgeometer".

No. 9, 1960

G. Weneki. A quartz mirror end gauge for an accurate parallel placing of measuring surfaces of the endpiece and the table of an instrument for linear measurements by means of interference fringes of equal thickness.

Proceedings of International Congress and Exhibition on Automation and Measurement Techniques "Interkama", held in October, 1960 in Dusseldorf.

No. 10, 1960

W. Bachman. Manometer with an elastic sensing element and the limiting possibilities of its utilization.

S. Herman. Balances in a nonuniform gravitational field. The article deals with accounting in precision measurements for the nonuniformity of the gravitational field and with gravitational curves and their effect on balances.

M. Schiska. Two-position regulator with an electronic master device and a transistorized amplifier for electrical measuring instruments.

Hire of measuring instruments in the USA. The hiring stock amounts to 13,000instruments of 70 different types.

ISA JOURNAL

No. 9, 1960

A new highly sensitive oscillator-transducer suitable for use in explosive premises. The transducer registers capacity changes of the order of 0.0001-0.00001 $\mu\mu$f.

No. 10, 1960

A. Schuler. The concept of a system of units of mass, weight, force, pressure and acceleration.

Terminology in vacuum techniques (in the sphere of instruments and measurements).

G. Dalke and V. Velkovich. A new ultrasonic commercial flowmeter for measuring liquid flow in units of mass and volume.

A. Votring and T. Macaveni. Application of an ultrasonic viscosimeter for determining concentrations in polymer solutions.

No. 11, 1960

B. Murray. A flowmeter with square root extraction by means of vibrating string.

No. 12, 1960

J. Cohen. Measuring flow by means of computers.

S. Akerman and J. Lord. Automatic photoelectric monochromatic pyrometer. Basic principles and a description of its construction are given.

A. Nylender. Nomogram for determining dynamic properties of moving-coil instruments.

Contemporary automatic regulations and control techniques in the USSR.

Informative conference on instruments and automation organized by the Instrument Society of America in New York, in September, 1960.

J. Johnston. Can the USA compete with the USSR in automation? Review of the contemporary automatic regulation and control techniques in the USSR from impressions gained by a delegate to the 1st Congress of the PFAC in Moscow.

REVUE DE MÉTROLOGIE

No. 6, 1960

Decree on the establishment of a higher school of metrology in France.

M. Ambar. Metrological problems in calibrating and using measured tanks, part I.

No. 7, 1960

M. Ambar. Metrological problems in calibrating and using measured tanks, part II.

No. 8, 1960

Standardization of measuring rods in measured tanks.

MICROTECNIC

No. 5, 1960

S. Bruce. Interference comparator for checking measures of length. The instrument has an additional contact mirror which rests on the measures being compared.

INSTRUMENT PRACTICE
AUTOMATION AND ELECTRONICS

No. 10, 1960

A new center for testing materials, instruments and equipment intended for public use.

V. Scholl. Nondestructive testing of materials by ultrasonic methods, part I.

No. 11, 1960

V. Scholl. Nondestructive testing of materials by ultrasonic methods, part II.

Symposium on measuring liquids in covered channels. Brief content of the papers read at the symposium held in Glasgow in September, 1960.

Special equipment for automatic measurement and recording of temperatures at many points in the production process.

No. 12, 1960

J. Vronhoven and A. Muelbaum. Electronic digital voltmeter. Description and circuit of the voltmeter, measuring range 0-1 v, error 1%.

COMMITTEE OF STANDARDS, MEASURES AND MEASURING INSTRUMENTS

NEW GOST (ALL-UNION STATE STANDARD) "MECHANICAL UNITS"

Translated from Izmeritel'naya Tekhnika, No. 3,
p. 64, March, 1961

On February 3, 1961, the Committee of Standards, Measures and Measuring Instruments approved at its meeting a new GOST "Mechanical Units", which will come into force on July 1, 1961. Contrary to the existing GOST 7664-55, the standard includes the new definitions of the meter and the second which were adopted in October, 1960 by the XI General Conference on Weights and Measures. It also contains denominations for all the derived units which were absent from the previous GOST. The standard recommends the MKS as the preferred system (which is a part of the International System of Units, approved by the XI General Conference on Weights and Measures), but permits the use for mechanical measurements of the CGS and MKG-WTS systems, as well as certain nonsystem units.

I. NEW SPECIFICATIONS FOR MEASURES AND MEASURING INSTRUMENTS APPROVED BY THE COMMITTEE

New Standards (registered in September-December, 1960)

GOST 9545-60. Glass cylinders for areometers. Replacing GOST 10078-39. Effective date April 1, 1961.

Gost 3720-60. Differential manometers. Replacing GOST 3720-54. Effective date January 1, 1962.

New Instructions for Checking Measures and Measuring Instruments

Instruction 100-60 for checking block-gauges. General propositions. Effective from January 1, 1961.

II. MEASURES AND MEASURING INSTRUMENTS APPROVED BY THE COMMITTEE AS THE RESULT OF STATE TESTS AND PASSED FOR USE IN THE USSR

(Registered in December, 1960 and January, 1961).

Microwebermeters, trade mark M199, of the Leningrad Sovnarkhoz (Council of National Economy). State Register No. 1418-60.

Portable voltammeters, trade mark É-504, of the Kiev Sovnarkhoz. State Register No. 1419-60.

Portable recording voltammeters, trade mark N-372, of Krasnodar Sovnarkhoz, State Register No. 1420-60.

Portable recording voltammeters, trade mark N-384, of the Krasnodar Sovnarkhoz. State Register No. 1421-60.

Amperehour dc electricity meters M640 and M640U of the Leningrad Sovnarkhoz. State Register No. 1422-60.

Current transformers, trade mark TPShL-10 of the Sverdlovsk Sovnarkhoz. State Register No. 1423-60.

Rack-mounted ammeters and voltmeters, trade mark M330, of the Krasnodar Sovnarkhoz. State Register No. 1424-60.

Mobile platform scales with weights, trade mark VPG-500 (b), of the Krasnodar Sovnarkhoz. State Register No. 1425-60.

Truck weighbridges with a dial indicator, trade mark ATs-25, of the Krasnodar Sovnarkhoz. State Register No. 1426-60.

5-ton trolley weighbridges with a dial indicator, trade mark VGTs-5, of the North Kazakhstan Sovnarkhoz. State Register No. 1427-60.

500 kg mobile platform scales with a dial indicator, trade mark VTsP-500, of the North Kazakhstan Sovnarkhoz. State Register No. 1428-60.

Microcolorimeters, trade mark KOL-52, of the Moscow Region Sovnarkhoz. State Register No. 1429-60.

Gas analyzers, trade mark GOU-1, of the Moscow Region Sovnarkhoz. State Register No. 1430-60.

Measuring lines, trade mark IKL-112, of the Moscow Region Sovnarkhoz. State Register No. 1431-60.

Voltmeter, trade mark S-50, of the Kiev Sovnarkhoz. State Register No. 1432-61.

Thermoelectric ammeters, trade mark T14, of the Leningrad Sovnarkhoz. State Register No. 1433-61.

Thermoelectric milliammeters, trade mark T15, of the Leningrad Sovnarkhoz. State Register No. 1434-61.

Voltammeters, trade mark M231, of the Omsk Sovnarkhoz. State Register No. 1435-61.

Magnetic field-strength meters, trade mark IMI-2, of Leningrad Sovnarkhoz. State Register No. 1436-61.

Infrasonic measuring oscillator, trade mark GNCh-1, of the Leningrad Sovnarkhoz. State Register No. 1437-61.

Stepped attenuators, trade mark AS-1 (D2-4), of the Leningrad Sovnarkhoz. State Register No. 1439-61.

Test set for low-power junction transistors, trade mark IPT-1 (L2-1), of the Belorussian Sovnarkhoz. State Register No. 1440-61.

Wide-band wavemeter, trade mark ShGV-S (Ch4-5), of the Ryazan' Sovnarkhoz. State Register No. 1441-61.

Platinrhodium-platinum reference thermocouples, trade mark PPO, of the Sverdlovsk Sovnarkhoz. State Register No. 1442-61.

Grade 3 reference weights of 0.5, 1 and 2 tons, of the Odessa Sovnarkhoz. State Register No. 811-61. Combined with Grade 3 reference weights previously entered in State Register No. 811.

Soviet Journals Available in Cover-to-Cover Translation

ABBREVIATION	RUSSIAN TITLE	TITLE OF TRANSLATION	PUBLISHER	TRANSLATION BEGAN Vol.	Issue	Year
AÉ	Atomnaya énergiya	Soviet Journal of Atomic Energy	Consultants Bureau	1	1	1956
Akust. zh.	Akusticheskii zhurnal	Soviet Physics – Acoustics	American Institute of Physics	1	1	1955
	Antibiotiki	Antibiotics	Consultants Bureau	4	1	1959
Astr(on). zh(urn).	Astronomicheskii zhurnal	Soviet Astronomy—AJ	American Institute of Physics	34	1	1957
Avto(mat). svarka	Avtomaticheskaya svarka	Automatic Welding	British Welding Research Association (London)		1	1959
	Avtomatika i Telemekhanika	Automation and Remote Control	Instrument Society of America	27	1	1956
	Biofizika	Biophysics	National Institutes of Health*		1	1957
	Biokhimiya	Biochemistry	Consultants Bureau	21	1	1956
Byull. éksp(erim). biol. i med.	Byulleten' éksperimental'noi biologii i meditsiny	Bulletin of Experimental Biology and Medicine	Consultants Bureau	41	1	1959
DAN (SSSR) Dokl(ady) AN SSSR	Doklady Akademii Nauk SSSR	The translation of this journal is published in sections, as follows:				
	Life Sciences	Doklady Biochemistry Section	American Institute of Biological Sciences	106	1	1956
		Doklady Biological Sciences Sections (Includes: Anatomy, biophysics, cytology, ecology, embryology, endocrinology, evolutionary morphology, genetics, histology, hydrobiology microbiology, morphology, parasitology, physiology, zoology sections)	American Institute of Biological Sciences	112	1	1957
		Doklady Botanical Sciences Sections (Includes: Botany, phytopathology, plant anatomy, plant ecology, plant embryology, plant physiology, plant morphology sections)	American Institute of Biological Sciences	112	1	1957
	Chemical Sciences	Proceedings of the Academy of Sciences of the USSR, Section: Chemical Technology	Consultants Bureau	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Section: Chemistry	Consultants Bureau	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Section: Physical Chemistry	Consultants Bureau	112	1	1957
	Earth Sciences	Doklady Earth Sciences Sections (Includes: Geochemistry, geology, geophysics, hydrogeology, mineralogy, paleontology, petrography, permafrost sections)	American Geological Institute	124	1	1959
		Proceedings of the Academy of Sciences of the USSR, Section: Geochemistry	Consultants Bureau	106-123	1 6	1957-1958
		Proceedings of the Academy of Sciences of the USSR, Sections: Geology	Consultants Bureau	106-123	1 6	1957-1958
	Mathematics	Doklady Soviet Mathematics	The American Mathematics Society	131	1	1961
	Physics	Soviet Physics—Doklady (Includes: Aerodynamics, astronomy, crystallography, cybernetics and control theory, electrical engineering, energetics, fluid mechanics, heat engineering, hydraulics, mathematical physics, mechanics, physics, technical physics, theory of elasticity sections)	American Institute of Physics	106	1	1956
		Proceedings of the Academy of Sciences of the USSR, Applied Physics Sections (does not include mathematical physics or physics sections)	Consultants Bureau	106-117	1	1956-1957
Derevoobrabat. prom-st'.	Derevoobrabatyvayushchaya promyshlennost'	Wood Processing Industry	Timber Development Association (London)		9	1959
	Élecktrosvyaz	Telecommunications	Massachusetts Institute of Technology*		1	1957
Éntom(ol). oboz(renie)	Éntomologicheskoe obozrenie	Entomological Review	American Institute of Biological Sciences	38	1	1959
Farmakol. (i) toksikol(ogiya)	Farmakologiya i toksikologiya	Pharmacology and Toxicology	Consultants Bureau	20	1	1957
FMM	Fizika metallov i metallovedenie	Physics of Metals and Metallography	Acta Metallurgica*	5	1	1957
Fiziol. zhurn. SSSR (im. Sechenova)	Fiziologicheskii zhurnal im. I. M. Sechenova	Sechenov Physiological Journal USSR	National Institutes of Health*		1	1957
Fiziol(ogiya) rast.	Fiziologiya rastenii	Plant Physiology	American Institute of Biological Sciences	4	1	1957
	Geokhimiya	Geochemistry	The Geochemical Society		1	1958
FTT	Fizika tverdogo tela	Soviet Physics—Solid State	American Institute of Physics	1	1	1959
Izmerit. tekh(nika)	Izmeritel'naya tekhnika	Measurement Techniques	Instrument Society of America		1	1959
Izv. AN SSSR, O(td). Kh(im). N(auk)	Izvestiya Akademii Nauk SSSR: Otdelenie khimicheskikh nauk	Bulletin of the Academy of Sciences of the USSR: Division of Chemical Sciences	Consultants Bureau		1	1952

Izmerit. tekh(nika)	Izmeritel'naya tekhnika	Measurement Techniques	Instrument Society of America		1	1959
Izv. AN SSSR, O(td). Kh(im). N(auk)	Izvestiya Akademii Nauk SSSR: Otdelenie khimicheskikh nauk	Bulletin of the Academy of Sciences of the USSR: Division of Chemical Sciences	Consultants Bureau		1	1952

continued

Izv. AN SSSR, O(td). T(ekhn). N(auk): Met(all). i top.	(see Met. i top.)					
Izv. AN SSSR Ser. fiz(ich).	Izvestiya Akademii Nauk SSSR: Seriya fizicheskaya	Bulletin of the Academy of Sciences of the USSR: Physical Series	Columbia Technical Translations		1	1954
Izv. AN SSSR Ser. geofiz.	Izvestiya Akademii Nauk SSSR: Seriya geofizicheskaya	Bulletin (Izvestiya) of the Academy of Sciences USSR: Geophysics Series	American Geophysical Union		1	1954
Izv. AN SSSR Ser. geol.	Izvestiya Akademii Nauk SSSR: Seriya geologicheskaya	Izvestiya of the Academy of Sciences of the USSR: Geologic Series	American Geological Institute		1	1958
Kauch. i rez.	Kauchuk i rezina	Soviet Rubber Technology	Research Association of British Rubber Manufacturers	18	1	1959
	Kinetika i kataliz	Kinetics and Catalysis	Consultants Bureau	1	3	1960
	Koks i khimiya	Coke and Chemistry USSR	Coal Tar Research Association (Leeds, England)		1	1958
Kolloidn. zh(urn).	Kolloidnyi zhurnal	Colloid Journal	Consultants Bureau	14	1	1952
	Kristallografiya	Soviet Physics – Crystallography	American Institute of Physics	2	1	1957
Metalov. i term. obrabot. metal.	Metallovedenie i termicheskaya obrabotka metallov	Metal Science and Heat Treatment of Metals	Acta Metallurgica	6	1	1958
	Metallurg	Metallurgist	Acta Metallurgica		1	1957
Met. i top.	Metallurgiya i topliva	Russian Metallurgy and Fuels	Eagle Technical Publications		1	1960
Mikrobiol.	Mikrobiologiya	Microbiology	American Institute of Biological Sciences	26	1	1957
OS	Optika i spektroskopiya	Optics and Spectroscopy	American Institute of Physics		6	1959
	Pochvovedenie	Soviet Soil Science	American Institute of Biological Sciences		1	1958
	Priborostroenie	Instrument Construction	British Scientific Instrument Research Association		1	1959
Pribory i tekhn. éks(perimenta)	Pribory i tekhnika éksperimenta	Instruments and Experimental Techniques	Instrument Society of America		1	1957
Prikl. matem. i mekh.	Prikladnaya matematika i mekhanika	Applied Mathematics and Mechanics	American Society of Mechanical Engineers		1	1958
PTÉ	(see Pribory i tekhn. éks.)					
	Problemy Severa	Problems of the North	National Research Council of Canada			
Radiotekh.	Radiotekhnika	Radio Engineering	Massachusetts Institute of Technology*	12	1	1957
Radiotekh. i élektronika	Radiotekhnika i élektronika	Radio Engineering and Electronics	Massachusetts Institute of Technology*	2	1	1957
	Stanki i instrument	Machines and Tooling	Production Engineering Research Assoc.		1	1959
	Stal'	Stal (In English)	Iron and Steel Institute		1	1959
Stek. i keram.	Steklo i keramika	Glass and Ceramics	Consultants Bureau	13	1	1956
Svaroch. proiz-vo	Svarochnoe proizvodstvo	Welding Production	British Welding Research Association		4	1959
Teor. veroyat. i prim.	Teoriya veroyatnostei i ee primenenie	Theory of Probability and Its Applications	Society for Industrial and Applied Mathematics		1	1956
Tsvet. Metally	Tsvetnye metally	Nonferrous Metals	Primary Sources		1	1960
UFN	Uspekhi fizicheskikh Nauk	Soviet Physics – Uspekhi (partial translation)	American Institute of Physics	66	1	1958
UKh	Uspekhi khimii	Russian Chemical Reviews	The Chemical Society (London)		1	1960
UMN	Uspekhi matematicheskikh nauk	Russian Mathematical Surveys	London Mathematical Society	15	1	1960
Usp. fiz. nauk	(see UFN)					
Usp. khim(ii)	(see UKh)					
Usp. matem. nauk	(see UMN)					
Usp. sovr. biol.	Uspekhi sovremennoi biologii	Russian Review of Biology	Oliver and Boyd		48	1959
Vest. mashinostroeniya	Vestnik mashinostroeniya	Russian Engineering Journal	Production Engineering Research Assoc.		4	1959
Vop. gem. i per. krovi	Voprosy gematologii i perelivaniya krovi	Problems of Hematology and Blood Transfusion	National Institutes of Health*		1	1957
Vop. onk.	Voprosy onkologii	Problems of Oncology	National Institutes of Health*		1	1957
Vop. virusol.	Voprosy virusologii	Problems of Virology	National Institutes of Health*		1	1957
Zav(odsk). lab(oratoriya)	Zavodskaya laboratoriya	Industrial Laboratory	Instrument Society of America	25	1	1959
ZhAKh Zh. anal(it). khimii	Zhurnal analiticheskoi khimii	Journal of Analytical Chemistry USSR	Consultants Bureau	7	1	1952
ZhETF Zh. éksperim. i teor. fiz.	Zhurnal éksperimental'noi i teoreticheskoi fiziki	Soviet Physics—JETP	American Institute of Physics	28	1	1955
ZhFKh Zh. fiz. khimii	Zhurnal fizicheskoi khimii	Russian Journal of Physical Chemistry	The Chemical Society (London)		7	1959
ZhMÉI Zh(urn). mikrobiol. épidemiol. i immunobiol.	Zhurnal mikrobiologii, épidemiologii i immunobiologii	Journal of Microbiology, Epidemiology and Immunobiology	National Institutes of Health*		1	1957
ZhNKh Zh(urn). neorgan(ich). khim(ii)	Zhurnal neorganicheskoi khimii	The Russian Journal of Inorganic Chemistry	The Chemical Society (London)		1	1959
ZhOKh Zh(urn). obshch(ei) khimii	Zhurnal obshchei khimii	Journal of General Chemistry USSR	Consultants Bureau	19	1	1949
ZhPKh Zh(urn). prikl. khimii	Zhurnal prikladnoi khimii	Journal of Applied Chemistry USSR	Consultants Bureau	23	1	1950
ZhSKh Zh(urn). strukt. khimii	Zhurnal strukturnoi khimii	Journal of Structural Chemistry	Consultants Bureau	1	1	1960
ZhTF Zh(urn). tekhn. fiz.	Zhurnal teknicheskoi fiziki	Soviet Physics—Technical Physics	American Institute of Physics	26	1	1956
Zh(urn). vyssh. nervn. deyat. (im. Pavlova)	Zhurnal vysshei nervnoi deyatel'nosti (im. I. P. Pavlova)	Pavlov Journal of Higher Nervous Activity	National Institutes of Health*		1	1958

*Sponsoring organization. Translation through 1960 issues is a publication of Pergamon Press.

SIGNIFICANCE OF ABBREVIATIONS MOST FREQUENTLY ENCOUNTERED IN SOVIET TECHNICAL PERIODICALS

AN SSSR	*Academy of Sciences, USSR*
FIAN	*Physics Institute, Academy of Sciences USSR*
GITI	*State Scientific and Technical Press*
GITTL	*State Press for Technical and Theoretical Literature*
GOI	*State Optical Institute*
GONTI	*State United Scientific and Technical Press*
Gosénergoizdat	*State Power Press*
Gosfizkhimizdat	*State Physical Chemistry Press*
Goskhimizdat	*State Chemistry Press*
GOST	*All-Union State Standard*
Gostekhizdat	*State Technical Press*
GTTI	*State Technical and Theoretical Press*
IAT	*Institute of Automation and Remote Control*
IF KhI	*Institute of Physical Chemistry Research*
IFP	*Institute of Physical Problems*
IL	*Foreign Literature Press*
IPF	*Institute of Applied Physics*
IPM	*Institute of Applied Mathematics*
IREA	*Institute of Chemical Reagents*
ISN (Izd. Sov. Nauk)	*Soviet Science Press*
IYap	*Institute of Nuclear Studies*
Izd	*Press (publishing house)*
LÉTI	*Leningrad Electrotechnical Institute*
LFTI	*Leningrad Institute of Physics and Technology*
LIM	*Leningrad Institute of Metals*
LITMiO	*Leningrad Institute of Precision Instruments and Optics*
Mashgiz	*State Scientific-Technical Press for Machine Construction Literature*
MGU	*Moscow State University*
Metallurgizdat	*Metallurgy Press*
MOPI	*Moscow Regional Pedagogical Institute*
NIAFIZ	*Scientific Research Association for Physics*
NIFI	*Scientific Research Institute of Physics*
NIIMM	*Scientific Research Institute of Mathematics and Mechanics*
NIKFI	*Scientific Institute of Motion Picture Photography*
NKTM	*People's Commissariat of the Heavy Machinery Industry*
Obrongiz	*State Press of the Defense Industry*
OIYaI	*Joint Institute of Nuclear Studies*
ONTI	*United Scientific and Technical Press*
OTI	*Division of Technical Information*
OTN	*Division of Technical Science*
RIAN	*Radium Institute, Academy of Sciences of the USSR*
SPB	*All-Union Special Planning Office*
Stroiizdat	*Construction Press*
URALFTI	*Ural Institute of Physics and Technology*
TsNIITMASh	*Central Scientific Research Institute of Technology and Machinery*
VNIIM	*All-Union Scientific Research Institute of Metrology*

NOTE: Abbreviations not on this list and not explained in the translation have been transliterated, no further information about their significance being available to us — *Publisher.*

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Russian original published by the Committee of Standards, Measures, and Measuring Instruments of the Council of Ministers, USSR. The articles in this journal are of interest to all who are engaged in the study and application of fundamental measurements. Both 1958 (bimonthly) and 1959-1961 (monthlies) available.

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INSTRUMENTS AND EXPERIMENTAL TECHNIQUES
Pribory i Tekhnika Éksperimenta

Russian original published by the Academy of Sciences, USSR. The articles in this journal relate to the function, construction, application, and operation of instruments in various fields of experimentation. 1958-1961 issues available.

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Russian original published by the Institute of Automation and Remote Control of the Academy of Sciences, USSR. The articles are concerned with analysis of all phases of automatic control theory and techniques. 1957-1961 1959, and 1960

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INDUSTRIAL LABORATORY – *Zavodskaya Laboratoriya*

Russian original published by the Ministry of Light Metals, USSR. The articles in this journal relate to instrumentation for analytical chemistry and to physical and mechanical methods of materials research and testing. 1958-1961 issues available.

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